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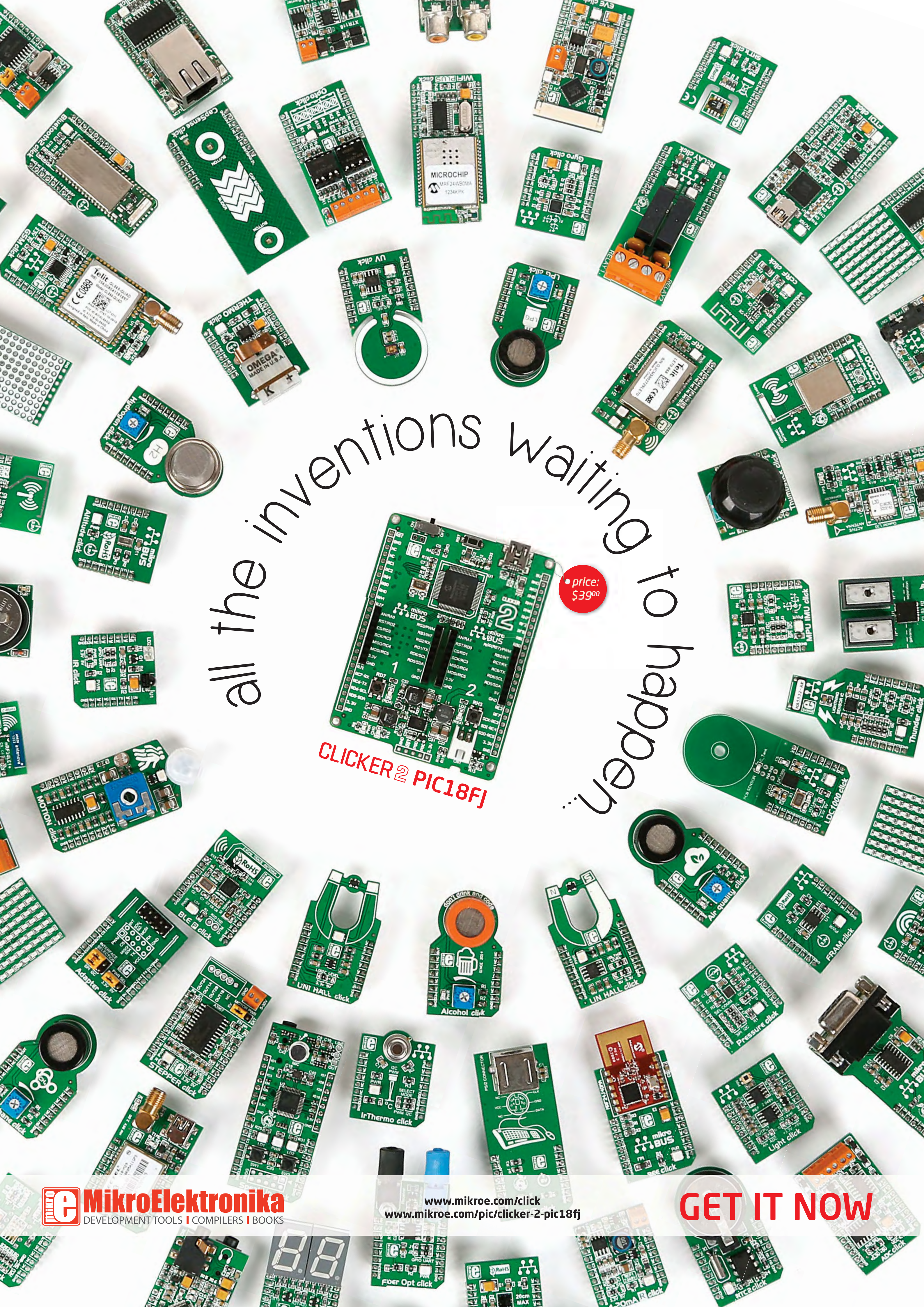
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Our April 2015 issue will be published on Thursday 5 March 2015, see page 72 for details.

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What's in store for 2015?

It's that time of year when policy makers, trend setters and crystal ball gazers like to try and predict what the 'next big thing' will be – and presumably where we will be spending our hard-earned cash. Not content to settle for just one prediction, Juniper Research (**www.juniperresearch.com**) has published a report on their 'Top 10 Tech Trends for 2015'. To be honest, their definition of 'Tech' is a little narrow – it's mostly related to computers and software, but there's enough diversity and meat on their tech bones to make it an interesting if somewhat short read.

Much of what they see for 2015 has been bubbling under the radar for the last few years and does not come as much of a surprise if you keep an eye on where 'Tech' is headed. For example, near field communications, wearable electronics and civilian drones are on their list. However, what did catch my attention was list item number one – 'Securing your data'.

Digital security is nothing new, and for decades there has been an ongoing arms race between privacy advocates and state security services over what is the acceptable ratio of privacy to security. Whether or not you feel that a certain loss of privacy to the UK's GCHQ, the NSA in the US and similar bodies in other countries is a price worth paying for personal and online security is a complex and polarising issue, but it has been given much greater prominence over the last few years with well-publicised leaks, hacks and the revelation of massive, industrial-scale data collection and surveillance programmes. Throw into the mix the need for protection from criminals, terrorists, hostile spies and anyone else with malign intent and you have one of the great social and technological issues of the early 21st century.

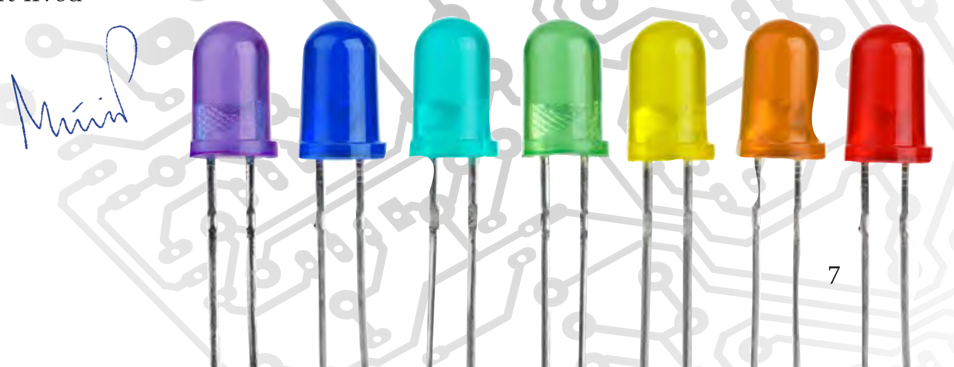
Stay secure...

Never slow to miss an opportunity, software and hardware companies are jumping into this space with an array of technologies that offer ways to protect online identity and personal data. Relevant buzzwords include encryption, tokenisation and biometric authentication – and at some point we may even have secure, 'unbreakable' systems based on quantum physics principles

... keep it simple

All of this is clever and technologically ingenious, but all too often the weakest link in a security system is not the technology, but the users – us! Like you, I have a myriad of accounts, from the trivial and superficial (little-used social media) to the vital (bank accounts, Internet commerce and of course *EPE* online!). They all need passwords and naturally I reuse and change all too rarely my collection of memorable alphanumeric strings. This leads to my one and only New Year's resolution for 2015 – yes, Juniper were correct to highlight cyber security – I resolve this year is to get serious about passwords. What does that actually mean? Well, I will finally get around to implementing some excellent advice from *Net Work* columnist Alan Winstanley, and start to use a password management system like Lastpass or Roboform…. well, that's my intention!

Stay safe out there and may your passwords be secure, memorable – and short-lived



NEWS

A roundup of the latest Everyday News from the world of electronics

Those irritating laws of physics... report by Barry Fox

A recent survey by chipmaker CSR found that only one in four music listeners are satisfied with the audio quality they are getting – because the music and audio industries have sacrificed quality for convenience.

Downloading and streaming has been facilitated by lossy compression systems, such as MP3, using inadequate, low bit rates. Playback through throwaway earbuds and underpowered, distortion-adding amp and speaker systems is now the norm.

Master Quality Authenticated

British Hi-Fi company Meridian recently announced MQA – Master Quality Authenticated – a new way of delivering high-quality audio at low bit rates. Says Bob Stuart, Meridian co-founder, 'I can't think of any other industry where quality has gone downhill.'

Too true – when I recently tried to buy a CD of music from Woody Allen movies, I could not find a physical copy anywhere and ended up purchasing an MP3 download from Amazon, which has a bit rate of 180kbps, even though the source CD has 44.1kHz/16-bit quality and MP3 can be used at 320kbps.

Wires, like discs, are now seen as old hat. Everyone is talking about home networking. It's easier to get high quality audio over Wi-Fi than Bluetooth, but Wi-Fi is power-hungry. So Bluetooth often has to be used.

Ignore the maths – just add magic!

Unfortunately, those wretched laws of physics – such as Shannon's theory (***see below***) on how much information a given bandwidth will carry – don't fit the industry's marketing hyperbole.

LG MUSIC flow system advert claiming 'true HD sound over Bluetooth' – what would Claude Shannon say?

The original, mandatory and default Bluetooth stereo standard is A2DP (advanced audio distribution profile) with SBC (sub-band coding). This is similar to early MPEG audio, and so old the patents have expired. The maximum sampling frequency of 48kHz and maximum bit rate of 345kbps are tailored to the bandwidth available from a Bluetooth

Claude Shannon – 'father of the information age'

Some people achieve much and are rightly famous, others make just as big an impression and almost no one has heard of them. Claude Shannon is very much a member of the latter category. The fact that he was an engineer may explain his modest level of fame.

So, what's the big deal with Mr Shannon? Put simply, he made the mathematical connections between bandwidth and the rate at which 'information' could be sent along a 'communication channel'. A channel could be a satellite link, copper wire, fibre optic cable or Bluetooth connection. This may sound like a very modern concept, a product of the Internet and mobile phone age, but his key findings were established in the 1940s when he was working on early cryptography and communications.

Shannon showed – precisely – how and why bandwidth mattered to the transmission of information, and crucially, the upper limit a particular bandwidth placed on transmission rates.

A flexible thinker with a sense of humour, he invented a flame-throwing trumpet and also that perennial favourite on YouTube, the 'Ultimate Machine'. This is a box with a single switch, which when pressed, causes a hand or finger to emerge from the box, reverse the switch and then disappear back into the box – you can see the fun at: **www.youtube.com/watch?v=cZ34RDn34Ws**

link. Quality is decidedly sub-CD, and the signal processing adds so much delay that lip sync with video is wrecked.

Most Bluetooth systems can now also use more efficient aptX stereo encoding. If devices have aptX they use it; if not, they fall back on A2DP.

aptX started out as a broadcast link system but has been tailored to fit into Bluetooth bandwidth. There is less latency delay because aptX relies on ADPCM (adaptive differential pulse code modulation) samples rather than data frames. This greatly reduces video lip sync problems.

CSR, which now owns aptX, claims 'CD performance' and 'full 'wired' audio quality'. In hard numbers, 4:1 compression squeezes CD-quality audio (16 bit, 44.1kHz) into a 352kbps data stream, with claimed 10Hz to 22kHz frequency response and 92dB dynamic range.

No one claims 'HD' quality, which is generally accepted to be better than CD quality, and usually 24 bit/96kHz or 192kHz – well, almost no one.

Bandwidth quart into a pint pot

Korean electronics giant LG has recently been advertising a home networked audio system called 'MUSIC flow', with the claim that 'everything works seamlessly over Wi-Fi or Bluetooth, in true HD sound' – see advert.

On LG's website there is a 'Learn More' hover link: HD Play: Play from any music source – even large capacity 24 bit/192 kHz HD music files – without sound ever dropping.' – **www.lg.com/uk/smart-hi-fi-systems#**

Their MUSIC flow technical specification refers to 'Audio Sound Mode – 24 bit/192kHz Sampling': **www.lg.com/uk/speakers-sound-systems/lg-NP8740**

And the website promises: 'High Definition Sound. HD Music Playback. 24 bit/192kHz. Listen to your music in High Definition sound quality (24 bit/192kHz) and play high quality lossless compressed files such as FLAC and WAV made possible through LGs Hi-Fi DAC [digital-to-analogue converter].' See: **www.lg.com/uk/speakers-sound-systems/lg-NP8740** and: **www.lg.com/uk/smarthifiaudio**

CSR's Senior Director of Audio, Chris Havell, confirms the obvious: 'I don't believe that Bluetooth supports HD Audio'.

I asked LG to explain how their MUSIC flow system can defy Shannon *et al* and deliver 'true HD sound' over a Bluetooth link. So far there has been no enlightenment.

Quantum Leap for UK Imaging Industry

Cameras which use just a single pixel to see through smoke, imaging systems which can time light to see around corners and miniature structures to create earthquake warning systems – these are just some of the quantum technologies set to be brought to market by a new 'Quantum Imaging Hub', led by the University of Glasgow, which will receive £29m in government funding.

One University of Glasgow-led project uses cheap, simple single-pixel sensors to build sophisticated ultraviolet or infrared video images much more affordably and conveniently than has been possible before. The sensors could be used in applications such as visualising gas leaks, seeing clearly through smoke, or looking under skin for tumours.

A second project, also led by the University of Glasgow, will use springs ten times thinner than a human hair to image minute changes in gravity fields. The work will enable a range of applications, including finding landmines, tracking magma moving under volcanoes and monitoring oil reserves to maximise extraction.

A new camera development led by Heriot-Watt University, uses highly advanced photon-timing techniques to recognise objects *around* corners, as well as images *through* walls or opaque biological tissue.

Parallax sensors

Online component and systems vendor Parallax has updated their web store with some interesting sensors that could form the heart of many fun and useful projects.

eTape Liquid Level Sensor

Their solid-state eTape sensors make it easier to monitor the height of fluids – liquid or a fine powder – in a container. Placed vertically, eTape senses the hydrostatic pressure applied by the fluid to the portion of the sensor that is submerged.

eTape sensors output a value in a continuous range corresponding to the fluid level along the entire length of the sensor, ideal where a single-point fluid level detection is not enough. The slim, low-profile sensor has no moving parts and fits unobtrusively where there may not be enough room for mechanical float devices.

eTape Liquid Level Sensors produce a variable resistance output. If your application is more suited to monitoring a voltage output, you may want to use the sensor with the optional eTape 0-5VDC Linear Resistance to Voltage Module. See: **www.parallax.com/product/29138**

PIR Mini Sensor

Need a small motion sensor? The PIR Mini Sensor's compact form factor is ideal for small-scale projects, and those where the electronics need to be hidden or protected. The narrow, vertically oriented board can slip through a hole in an enclosure to hide and protect the electronics, leaving only the small lens visible on the outside.

The PIR Mini Sensor detects motion up to 12 feet (3.65m) away. Like the other passive infrared sensors, the PIR Mini is a pyroelectric device that senses and responds to changes in the infrared levels emitted by nearby objects. See: **www.parallax.com/product/28033**

eTape Liquid Level Sensor (above) – PIR sensor (top right)

Astro Pi

Leading UK space organisations have joined forces with British ESA Astronaut Tim Peake and Raspberry Pi to offer students a chance to devise and code their own apps or experiment to run in space. Two Raspberry Pi computers are planned to be flown to the International Space Station as part of Tim's six-month mission, and both will be connected to a new 'Astro Pi' board, loaded with a host of sensors and gadgets. Further details and resources are available at: **www.raspberrypi.org/astro-pi**

Super Smooth, Full-range, 10A/230V Speed Controller for Universal Motors

by
John Clarke

Most mains motor speed controllers aren't very good! They often have very poor low-speed control or won't allow control right up to the motor's maximum speed – or both! Here's one that is exceptional: a microcontroller-powered full-wave circuit that overcomes both these problems and gives extremely smooth control as well. It's ideal for use with electric drills, lawn strimmers, circular saws, routers or any other appliance with universal (ie, brush-type) motors, rated up to 10A.

This 10A electronic speed controller provides an impressively smooth running universal motor that can be adjusted from very slow up to full speed. Using the feedback control, the motor can be set to maintain its speed even under load.

A similar *230VAC 10A Full-Wave Brush Motor Speed Controller* was published in June 2011. This controller worked well, but this latest controller has additional features which provide significant improvements. These include improvements to not only the core motor control, but also added protection to the controller circuitry, such as cycle-by-cycle over-current limiting and soft starting.

We do need to mention here that this controller is ***not*** suitable for use with induction motors, such as are typically used with compressors, bench grinders, lathes and pumps. For more information do make sure you read the panel entitled: 'What motors can be controlled?'

Why is it so good?

So, why is this controller so good at driving brush-type motors, particularly at slow running speeds and for full-speed operation? It's all to do with the type of voltage waveform that is used to provide speed control.

Typically, brush motor speed controllers use a simple phase-control circuit. Shortcomings of phase control are immediately apparent when using this design. One is that the maximum speed from the motor when under full-speed

control adjustment is significantly reduced – up to 25% or more – compared to running directly from the mains.

So, for an electric drill that normally runs at say 3000 RPM, the maximum speed might be reduced to around 2200 RPM. This is inevitable with a controller circuit that effectively half-wave rectifies the 230VAC mains waveform to give a maximum output voltage of around 162V RMS.

The second drawback of a phase control design has to do with low-speed control. While such circuits often do allow your drill or other appliance to run at quite low speeds, the result is that there isn't much torque available and the speed regulation is poor.

This means that if you're operating the drill at a low speed and you put a reasonable load on it, its speed will drop right away or it may stall completely.

Worse still, the motor will tend to 'cog'. Cogging is caused by erratic firing of the main switching device (a triac) within the drill speed controller, so that the motor receives intermittent bursts of power.

An electric motor that is cogging badly is virtually useless, and the only cure is to increase the speed setting, defeating the purpose of a speed controller if you want to operate at low speed.

What's the alternative?

Both of these drawbacks are basically eliminated with our new *Speed Controller for Universal Motors*.

The design does not use phase-control circuitry, but uses switch-mode power supply techniques to produce an outstanding controller for all types of universal brush motors. (Virtually all mains-powered [handheld] power tools and many appliances use universal motors. These are series-wound motors with brushes.)

It has very low-speed control with excellent maintenance of speed under load. Additionally, it will run the motor over its full speed range, even at full speed if required.

Most power tools will do a better job if they have a speed control. For example, electric drills should be slowed when using larger drill bits, as they make a cleaner cut.

Similarly, it is useful to be able to slow routers, jigsaws and even circular saws when cutting some materials, particularly plastics (many plastics actually melt and then meld if the speed is too high).

The same comments apply to sanding and polishing tools and even electric strimmers – they're less likely to snap their lines when slowed.

Phase control

Before we continue, we should explain what we mean by 'phase control' so that we can illustrate the benefits of this new design.

As you know, the mains (AC) voltage closely follows a sinewave. It starts at zero, rises to a peak, falls back to zero, then does the same thing in the opposite direction. This repeats at 50 times each second (50Hz).

A motor connected to the mains makes full use of the energy from each cycle so that it runs at its maximum speed. But if we were only to supply a portion of the waveform, with less energy available to power it, the motor would not run so fast.

By varying the time during each half cycle when power is applied, you would have a variable speed control.

This then is the basis of phase control: feed power very early in the cycle and it runs fast; delay power until much later in the cycle and it runs slowly.

The term 'phase control' comes about because the timing of the trigger pulses is varied with respect to the phase of the mains sinewave. Phase control has in the past been the basis for incandescent lamp dimmers and even heater controls. (By the way, phase control is not generally suitable for fluorescent and compact fluorescent lamps.)

The oscilloscope waveform of Fig.1 shows the chopped waveform from a phase-controlled circuit when a motor is driven at a fast speed. Fig. 2 shows the waveform from the phase-controlled speed control at a lower setting. At the low setting the motor has 45V RMS applied, while at the higher setting, the motor has 138V RMS applied to it

These examples show only the positive half of the mains waveform being used, as is the normal case with a phase-controlled circuit. This automatically limits the amount of energy which can be delivered to the motor – the power available from the negative waveform cycles is not used.

It means that in a half-wave phase-control circuit, the range of control is limited to a relatively small range of speeds.

For the motor to run at full speed, it would need to be fed with both the positive and negative half-cycles of the 50Hz mains waveform.

Features

- Extremely smooth and precise motor speed control
- Speed can be controlled from zero to maximum
- Superb speed regulation under load
- Adjustable speed regulation with feedback control
- Excellent low-speed motor operation
- 2300W (10A) rating
- Cycle-by-cycle current overload protection
- Over-current limiting
- Soft starting
- NTC Thermistor for initial surge current limiting
- Fused protection
- Rugged case with interference suppression included
- For 230VAC brush (universal) motors

Why build this when you can buy a power tool with inbuilt controller?

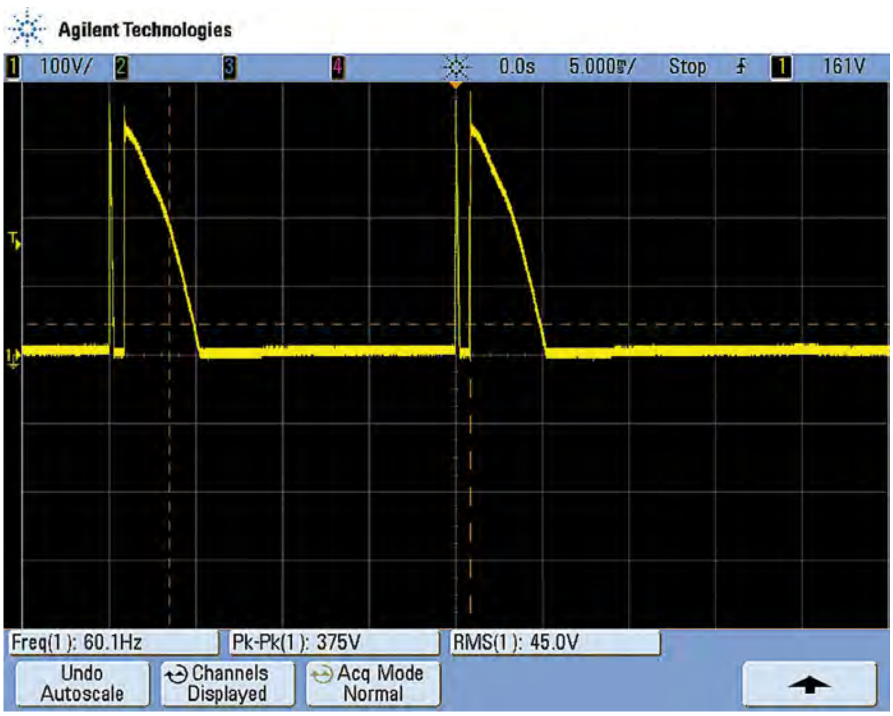
Many hand power tools these days have inbuilt (trigger or dial) speed controllers. And many cost less than this stand-alone controller kit. So why would you build this one?

Quite simply, this is better! That's no idle boast – everyone who has tried this out has been very pleasantly surprised. You won't believe how smooth the control is, nor how much 'grunt' you get at low speed. Or any of the other features this new 230V 10A Universal Motor Speed Controller offers! It's not just better than any previous controller – it's significantly better . . .

Or perhaps you have an existing favourite power tool that doesn't have speed control – build this and enjoy new versatility!

Fig.1. These waveforms illustrate the operation of a typical phase-controlled SCR when driving a typical electric drill. Here the SCR is triggered early in the positive half-cycle, so the motor voltage is 138V RMS and it runs at a relatively high speed. The motor can never run at maximum speed for another reason: half of the energy is unavailable because only one half of the cycle is used. (Even if the whole half-cycle could be fed to the motor, it could only ever be about 162V RMS). Also notice that there is considerable hash at the beginning of each positive half-cycle, caused by interaction between the drill's commutator and the triac.

Fig.2. Compare Fig.1 with the waveform below where the SCR is triggered much later in the half-cycle, meaning less power is available to the motor – the voltage being fed to the motor here is just 45V RMS. While it does run much slower – the aim of the exercise, of course – it suffers from low torque and is also liable to 'cog'. Note the frequency error in both these screen grabs, which is caused by hash on the waveform and the fact that the SCR triggering is erratic.



Normally this is not possible with a phase-control circuit that uses an SCR (silicon controlled rectifier), which is, effectively, a controlled diode that only conducts in one direction.

While a triac could be used to switch the full 50Hz mains for phase control (ie, both positive and negative-going half cycles), it is difficult to achieve and still incorporate constant speed control under load without a complex circuit.

Additionally, another big problem with conventional phase-controlled circuits is that the trigger pulse applied to the triac or SCR is very short.

If this corresponds with the instant when the brushes hit an open circuit portion of the commutator, no current will flow and the motor will miss out on a whole cycle of the mains waveform.

Similarly, even if the triac or SCR has been correctly triggered on, the SCR or triac may switch off again as current falls to zero when a brush passes an open circuit on the commutator.

This problem is more critical at low speed settings and is one of the reasons for the 'cogging' behaviour referred to earlier.

Incidentally, the sparks you see when you look into a universal (brush type) motor are mostly caused by brushes passing over the open circuit section of the commutator.

Typically, a power drill might have a dozen or more open circuit sections on the commutator. These open circuit sections or gaps in the commutator are necessary to keep motor windings separate.

Speed regulation

Most phase-controlled SCR or triac speed control circuits claim to include a form of feedback that is designed to maintain the speed of the motor under load.

They rely upon the fact that a motor can be used as a generator when it is spinning with no power applied. When the motor is loaded and the motor speed slows, the back-EMF (electromotive force) produced by the motor drops and the circuit compensates by triggering the SCR earlier in the mains cycle. This earlier triggering helps to drive the motor at the original speed.

In practice, however, the back-EMF generated by most series motors when the SCR or triac is not conducting is either very low or non-existent. This is due in part because there is no field current and the generation of voltage is

Parts List – Super Smooth, Full-range, 10A/230V Speed Controller for Universal Motors

1 PCB, code 10102141, 112 × 141mm, available from the *EPE PCB Service*
1 metal diecast case, 171 × 121 × 55mm
1 front panel label, 168 × 118mm
1 13A single switched mains power outlet
1 240VAC 10A PCB mount EMI filter (Schaffner FN 405-10-02 or equivalent)
1 NTC Thermistor (SL32 10015) (Element14 Cat.1653459)
1 10A IEC mains lead (3-pin mains plug to IEC line female connector)
1 IEC male chassis connector with fuse
1 10A M205 fast-blow fuse (F1)
2 knobs to suit potentiometer shafts
2 2-way PCB-mount screw terminal blocks, 5.08mm spacing (CON1)
5 6.35mm PCB-mount male spade connectors with 5.08mm pin spacing
5 6.35mm insulated female spade quick connectors with 4-8mm wire diameter entry
2 5.3mm ID insulated quick connect crimp eyelets with 4-6mm wire diameter entry
1 18-pin DIL IC socket
1 M4 × 10mm pan head or countersunk screw (Earth to case)
1 M4 × 10mm countersunk screw (Earth to lid)
2 M4 × 15mm pan head screws (power outlet mounting)
1 M4 × 20mm pan head screw (BR1 mounting)
5 M4 nuts
4 4mm star washers
2 M3 × 10mm countersunk screws (for IEC Connector)
2 M3 × 15mm pan head screws (for Q1 and D1)
8 M3 nuts
2 3mm star washers
2 M3.5 × 6mm screws (for mounting PCB to case)
4 stick-on rubber feet
6 100mm cable ties
2 TO-3P silicone insulating washers
1 400mm length of blue 10A mains wire
1 400mm length of brown 10A mains wire
1 400mm length of green/yellow 10A mains wire
1 200mm length of brown 7.5A main wire
1 200mm length of blue 7.5A mains wire
1 70mm length of black 5mm heatshrink tubing
1 10mm length of red 5mm heatshrink tubing
1 40mm length of 2.5mm Vidaflex heat resistant sleeving

Semiconductors

1 PIC16F88-I/P microcontroller programmed with 1010214A.hex (IC1)
1 LMC6482AIN dual CMOS op amp (IC2)
1 IR2125 PDIP current-limiting single channel MOSFET/IGBT driver (IC3)
1 LP2950ACZ-5 5V regulator (REG1)
1 STGW40N120KD 1200V 40A NPN IGBT (Q1) (Element14 Cat. 2344080)
1 2N7000 N-channel MOSFET (Q2)
1 STTH3012W 30A 1200V TO-247 ultra-fast recovery diode (D1) (Element14 Cat.1295262)
1 1N4148 general purpose diode (D2)
1 15V 1W Zener diode (ZD1)
1 35A 400V or 600V bridge rectifier (BR1) (PCB mount) or (with quick-connect terminals; with additional components required. See below*)
1 W04 400V 1.2A bridge rectifier (BR2)

Capacitors

2 100µF 16V PC electrolytic
5 1µF 50V monolithic multilayer (MMC)
1 470nF 63V or 100V MKT polyester
2 220nF 250VAC X2 class MKT polyester
1 100nF 250VAC X2 class MKT polyester
5 100nF 63V or 100V MKT polyester
1 15nF 63V or 100nF MKT polyester
1 10nF 250VAC X2 class MKT polyester
1 470pF ceramic

Resistors [0.25W 1%] #=1W, 5%

2 1MΩ#	1 1MΩ	3 10kΩ	1 4.7kΩ
2 2.2kΩ	1 1kΩ	2 470Ω#	1 330Ω
3 100Ω#	1 10Ω	1 4.7Ω 0.25W 5%	

2 24mm 10kΩ linear single gang potentiometers (VR1,VR2)
1 0.010Ω 3W low-ohm shunt resistor (TT Electronics, Wellwyn OAR3 R010)
1 10kΩ miniature trimpot (horizontal mount with 5mm pin spacing) (VR3)

*** Additional components**

4 6.35mm PCB mount male spade connectors with 5.08mm pin spacing
4 6.35mm insulated female spade quick connectors with 4-8mm wire diameter entry
1 80mm length of 10mm diameter heatshrink tubing

Jaycar Electronics produce a kit for this speed controller: Cat KC5526, priced at $149.00
See: www.jaycar.com.au/productView.asp?ID=KC5526.
(Note; the mains power outlet in the Jaycar kit will need to be changed for one suitable for the country in which the controller is used)

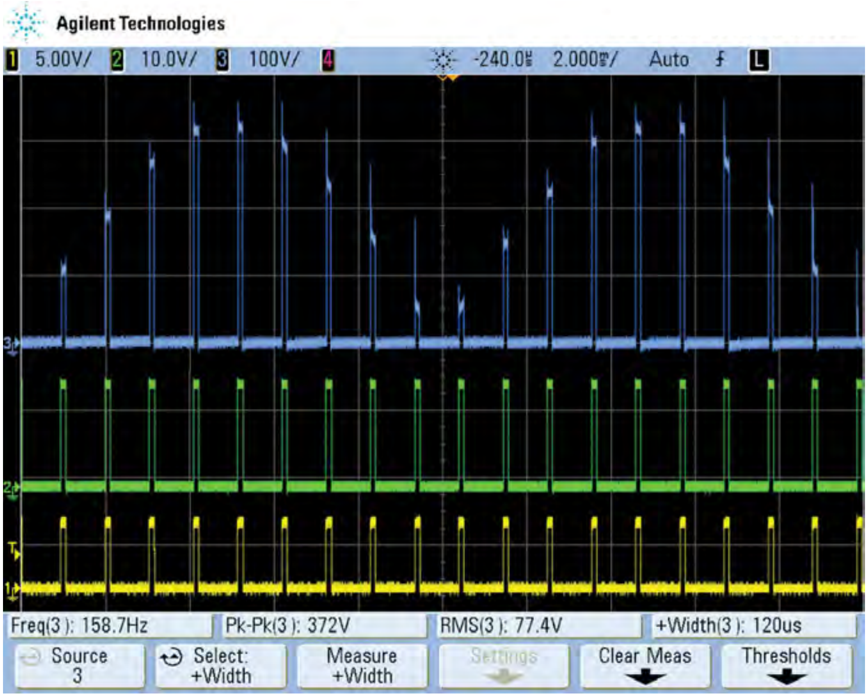


Fig.3. this series of scope screen grabs, taken with the controller driving a typical handyman electric saw, shows the voltage waveforms applied to the motor at progressively higher speed settings. This is the lowest setting, with very short pulses from the IGBT delivering just 77.4V RMS to the motor. The yellow trace shows the output from IC1 (as applied to the IGBT driver), while the green trace is the output from that driver. The top (blue) trace shows the voltage actually applied to the motor via the GPO. You can see that it follows the full-wave-rectified mains 'outline' but the pulses themselves are very narrow.

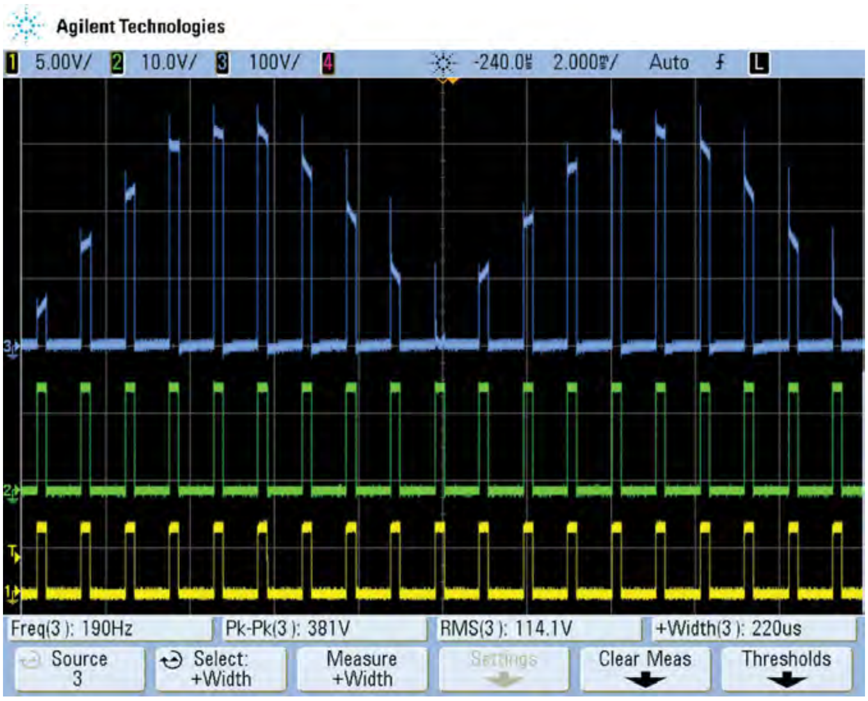


Fig.4 shows a significantly higher speed setting (114V RMS) with the IGBT being switched on with longer pulses. The yellow and green traces remain constant in their amplitude, but of course the pulses are wider, therefore delivering more energy. By the way, the spikes on the leading edges of the motor waveform (blue trace) mainly appear to be an artefact of the measurement method (ie, they are not actually present!).

only due to remanent magnetism in the motor core. If any back-EMF produced, it is too late after the end of each half-cycle to have a worthwhile effect on the circuit triggering in the next half-cycle.

So while phase control is simple and cheap, it is not an ideal method for controlling motor speed. Instead, we use a different method – as follows.

Pulse-width modulation

Our new speed-control circuit uses pulse-width modulation (PWM) and a different feedback method for speed regulation that effectively solves the above problems associated with phase control. Fig.3 to Fig.6 show the voltage waveforms applied to the motor at progressively higher settings from very low to full speed.

What happens is that we rectify the mains voltage and then chop it up at a switching rate of about 980Hz using a high voltage IGBT (insulated gate bipolar transistor). For the lowest speed setting (Fig.3), the pulses are very narrow and for the higher speed settings the pulses applied to the motor are progressively wider.

There are between 9 and 10 pulses during each half cycle, so the motor receives a more continuous stream of current compared to when driven via phase control. As a result, the motor operates very smoothly over the whole of its speed range.

For speed regulation, the circuit does not rely upon the back-EMF from the motor. Instead, it monitors the current through the motor and adjusts the pulse width to maintain the motor speed. When a motor is idling, it draws a certain amount of current to keep itself running. When the motor is loaded, the motor speed drops and the current drawn by the motor increases. The motor controller senses this and then compensates for this speed drop by widening the pulse width to maintain motor speed.

Block diagram

Fig.7 shows the basic circuit arrangement. The 230VAC input waveform is fed through a filter and full-wave rectified.

An NTC thermistor in series between the full-wave rectified supply and the motor limits the initial surge current drawn by the motor. The thermistor has a relatively high resistance when cold; as it heats up, the resistance drops allowing full power to be applied to

the motor when necessary. The NTC thermistor is ideal for use with heavy-current appliances to reduce the start up current.

The resulting positive-polarity waveform is fed to one side of the motor. The other motor terminal is switched on and off via IGBT Q1. An IGBT is a hybrid of a MOSFET and bipolar transistor. It has the high impedance gate drive of a MOSFET, but high current handling at high voltages, like a power transistor.

The IGBT we are using has a 40A, 1200V rating (120A peak) and can even withstand a short circuit for 10µs. Switching of the IGBT is under the control of the gate driver, IC3, which in turn is controlled by the microcontroller, IC1.

IC1 monitors the speed potentiometer VR1 and produces a PWM signal that is proportional to the speed setting. So for higher speed settings of VR1, the PWM output from IC1 will be wide pulses, while a lower speed setting will reduce the pulse width.

The PWM output is fed to IC3 that then drives the high voltage IGBT (Q1). Diode D1 is a fast-recovery type to conduct the motor current when Q1 is switched off.

The 'snubber' across Q1, consisting of a 33Ω resistor and 10nF capacitor, suppresses excessive voltage excursions.

The very low value resistor, R1, is included for monitoring current flow through the motor. This current measurement is used for two purposes. First, the current is monitored by IC3 and this IC will reduce drive to the IGBT should the current go beyond a peak of about 23A. This IC monitors the peak current during each switching cycle to protect the IGBT from damage due to over current.

For speed regulation, the voltage across R1 is filtered, sampled and amplified. Sampling of the current occurs only whenever Q1 is switched on to drive the motor. The current feedback is held at this sampled voltage level when the motor is switched off. The amplified current measurement is monitored by IC1 and averaged over a 10ms period, thus capturing a full half mains cycle of current.

An over-current comparator is included and is also monitored by IC1. It differs from the cycle-by-cycle, peak over-current protection provided by IC3. It works as follows: whenever the average current exceeds 15A, IC1 begins to reduce the duty cycle of the

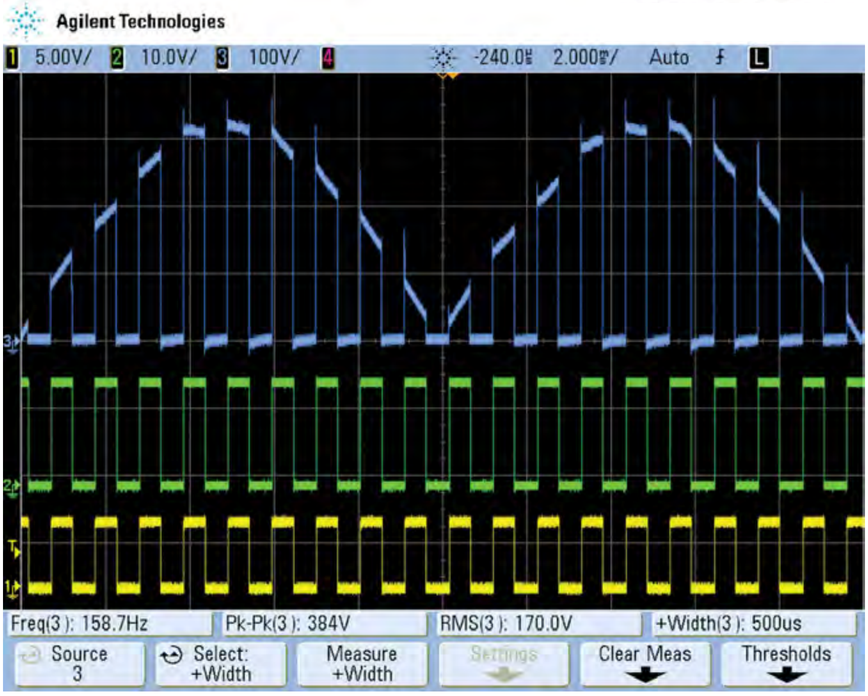


Fig.5. similarly, shows an even higher speed setting – very close to 50% duty cycle – with now 170V RMS being delivered to the motor by the IGBT. Motor speed would already be higher than that capable of a phase-controlled circuit and shows how good this circuit is.

Incidentally, all the waveforms displayed in this series of figures have been measured using high-voltage differential probes on the oscilloscope. Do not attempt to make any of the measurements using conventional probes and an isolating transformer – as there is a risk that you will blow the IGBT, the fast recovery diode, D1 and the gate driver chip, IC3. *We write this from bitter experience!*

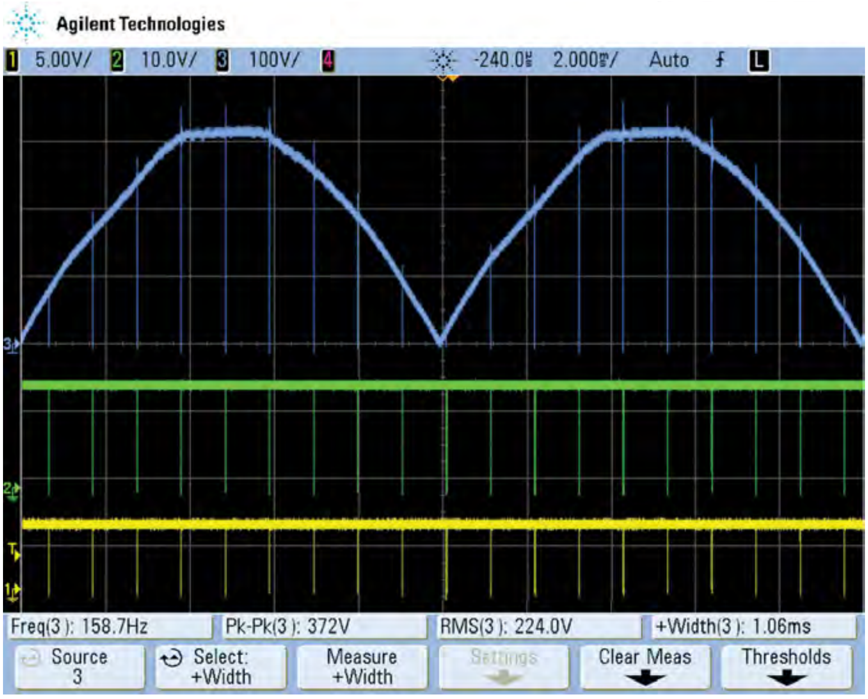


Fig.6. here the IGBT is virtually full-on, delivering maximum voltage to the motor. The drive pulses are virtually at 100% so the motor would be running at the same speed (or very close to it) just as if it were plugged directly into the 230V AC mains. However, the RMS voltage reads a little lower than expected, due to the fact that the spikes which were present in the earlier waveforms are no longer there to confuse the scope.

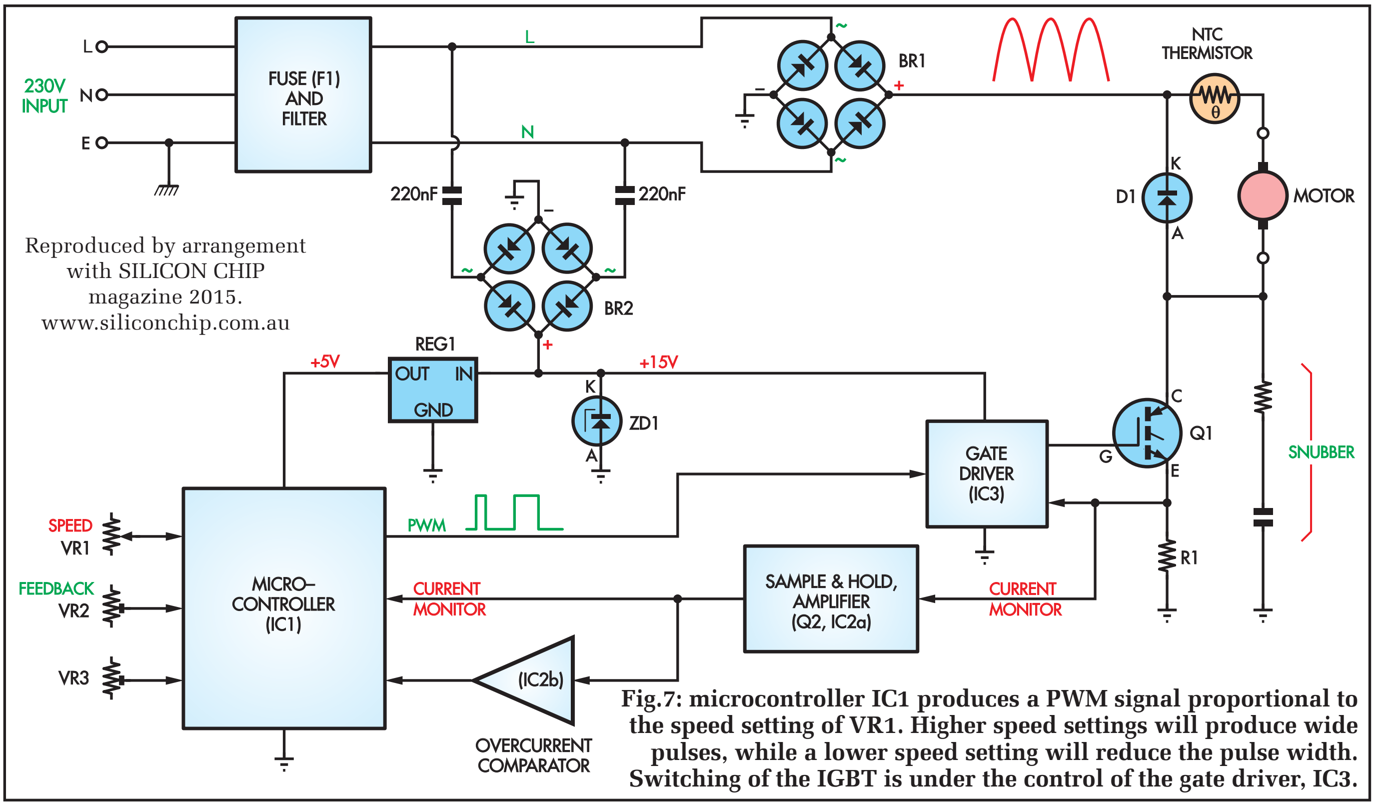


Fig.7: microcontroller IC1 produces a PWM signal proportional to the speed setting of VR1. Higher speed settings will produce wide pulses, while a lower speed setting will reduce the pulse width. Switching of the IGBT is under the control of the gate driver, IC3.

PWM drive until the comparator output switches low, indicating a lower current. It thus provides an overall current limit.

VR2 and VR3 are for the feedback control. VR2 is a potentiometer that's externally adjustable as it is mounted on the lid of the controller. Alternatively, if you prefer not to have VR2 mounted on the case lid, then VR3 can be used to set the degree of feedback. VR3 is a trimpot installed inside the controller. The feedback control adjusts by how much the duty cycle of the PWM motor drive is increased under load.

One of the advantages of using a microcontroller is that the feedback control can include features not possible with conventional circuitry.

First, when starting the motor from stopped, any feedback control is inactive until the motor reaches the speed that it is set to run at by the speed control. This motor-start operation can be activated by turning the speed control up (from fully anticlockwise) or by switching on the motor. The lack of feedback control prevents the motor giving a large overshoot in its speed when it first starts up. A stopped motor is detected each time the average motor current drops to zero.

Second, the microcontroller can 'dial out' the idle (no load) motor current, so motor speed is not increased markedly with increased feedback settings. If this is dialled out, only the extra current drawn by the motor under load is used by IC1 to adjust PWM to maintain motor speed. This feature is especially useful with higher-current motors.

The motor idle current is dialled out by running the motor at the speed required with the speed control and with the feedback control set to its minimum setting. The motor's idle current will then be recorded by IC1 and feedback will only operate when motor current exceeds this current.

Any changes that increase the motor speed, either through a change in position of the speed control or starting the motor, the PWM signal is varied at a slow rate with small increases made each 10ms. For a complete ramp-up in motor speed over the possible 255 speed settings, full PWM duty is only available after ramping up over 2.54s.

Circuit description

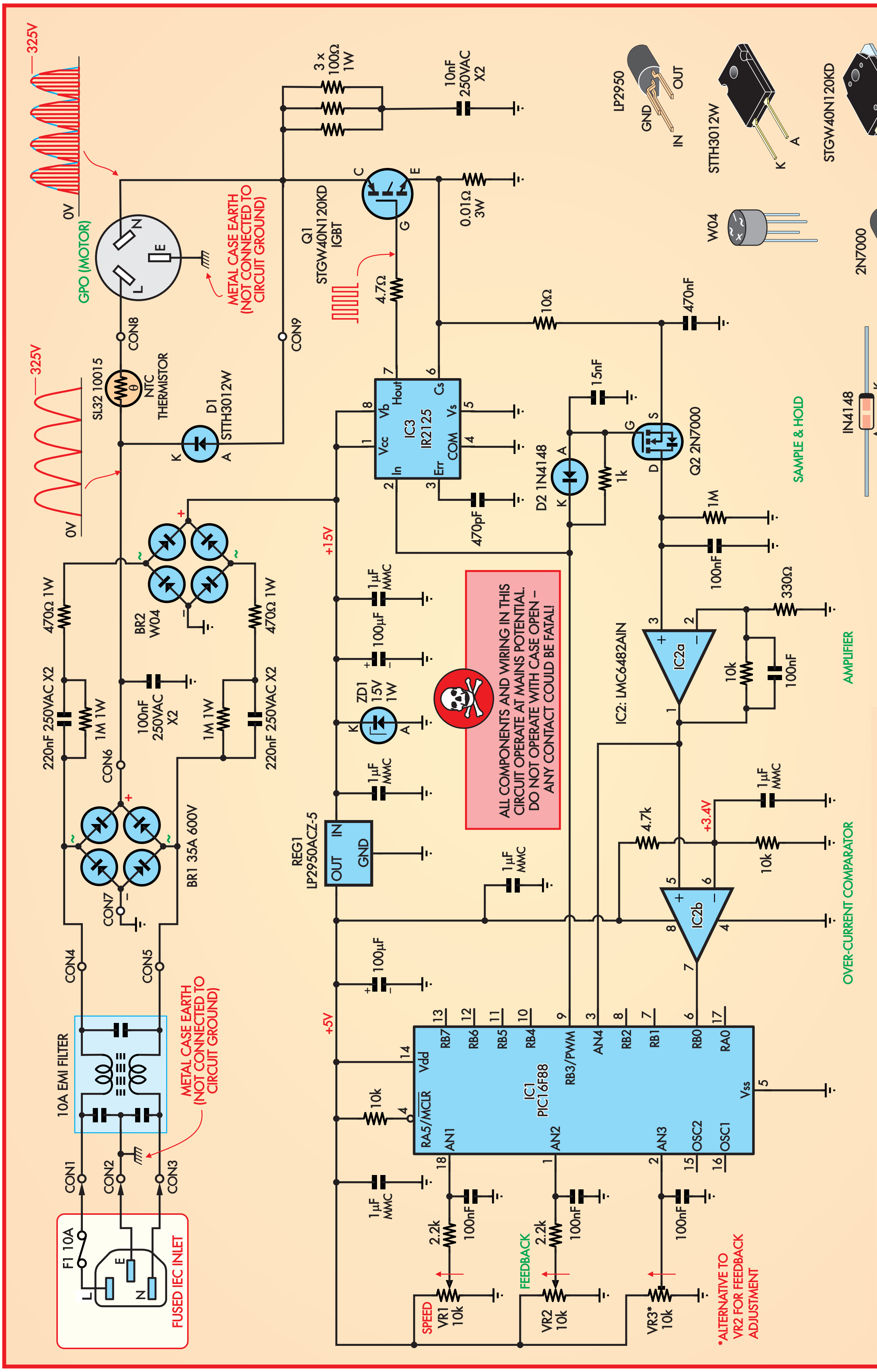
The circuit for the *Motor Speed Controller* is shown in Fig.8. It comprises three ICs, several diodes, resistors and capacitors, plus the high-voltage IGBT, Q1. Power for the circuit is derived directly from the 230VAC mains.

The entire circuit floats at mains potential and is therefore unsafe to touch whenever the circuit is connected to the mains. Also note that the circuit ground is floating at mains potential (it is not connected to mains earth, which connects only to the metal case). **The unit must be constructed in a rugged, earthed metal case.**

Mains power supplied to the *Controller* circuit is via a fuse, F1, that's integral to the IEC input connector. This fuse protects the circuit against excessive current flow that can occur with a short across the motor. An

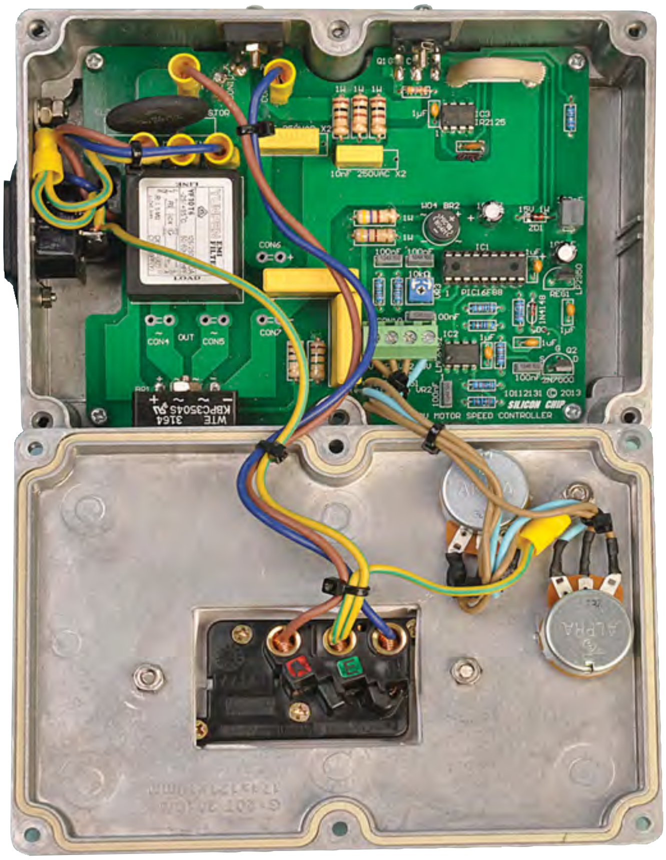
Specifications

Rating	**10A, 230VAC**
Speed adjustment	**Zero to motor's maximum**
PWM frequency	**980Hz**
Cycle-by-cycle current limiting	**23A peak**
Average current limiting	**15A**
Soft-start rate	**Up to 2.54s from zero to full speed**
NTC thermistor	**10Ω at 20°C, <0.1Ω @10A**



10A 230V AC INTELLIGENT SPEED CONTROLLER

Fig.8: the complete circuit diagram. Note the warning – it is not safe to work on an open case when connected to the 230V AC mains!



Ready for next month's construction details, here's the inside view of the new *Motor Speed Controller*. It's fully self-contained within a rugged, earthed diecast box.

electromagnetic interference (EMI) filter reduces switching artefacts from the IGBT and motor being radiated back to the mains wiring. This is a commercially-made filter that consists of a pair of 2.2nF to 3.3nF capacitors from mains live and neutral to earth, followed by a 0.3 to 0.4mA current-compensated series choke for each line, then a 15nF to 100nF capacitor across the load terminals (actual values depend on manufacturer).

BR1 is a 35A bridge rectifier with a 400V or 600V rating. The bridge provides the circuit with the positive full-wave rectified mains voltage to power the motor. This supply is filtered using a 100nF 250VAC capacitor. The capacitor does not provide a smoothed DC supply. Instead, the capacitor just filters out much of the high frequency switching noise on the supply due to the motor and also helps to reduce the voltage induced when the IGBT is switched off and D1 becomes forward biased.

A separate supply arrangement is used for the low-voltage circuitry. Instead of just using high-power resistors to limit current to a Zener diode, we use a capacitor-coupled separate bridge rectifier in order to reduce power and more importantly heat dissipation inside the controller case.

The second rectifier (BR2) is fed via two 220nF capacitors and series 470Ω resistors.

The 220nF capacitors are used to provide an impedance-limited current to the 15V Zener diode, ZD1. For 50Hz, the impedance of each 220nF capacitor is 14.5kΩ. This, plus the 470Ω limits the current through ZD1. A 100µF capacitor across the resulting 15V supply smooths the voltage to a near-constant value.

The 470Ω resistors in series with the 220nF capacitors are there to limit surge current when power is first applied to the circuit. The surge current could be high should power be switched on at the peak voltage of the mains waveform.

1MΩ resistors across the capacitors are there to discharge any stored voltage when the power is switched off.

Without these, the capacitor could have high voltage stored ready to provide an electric shock to anyone touching the capacitor when, for example, trouble shooting the circuit (even when 230V AC power is disconnected).

Low voltage supplies

The 15V supply powers the IGBT driver IC3 directly, while a low-power 5V regulator derived from the 15V line supplies both IC1 and IC2. The 100µF and 1µF capacitors at the regulator's output and input ensure the regulator remains stable and that it can provide transient current without losing regulation.

IC3 is a dedicated MOSFET (or IGBT) driver used as a low-side driver where the output produces a 15V gate drive with respect to the circuit ground.

Apart from providing gate drive for the IGBT, IC3 also protects the IGBT. It does this in several ways. First, the gate drive is a high-current pulse to minimise the time that the IGBT is in its unsaturated state to minimise power dissipation.

Second, current is monitored across a 0.01Ω resistance between the emitter and the circuit ground. Whenever the voltage across this resistor rises above 230mV, representing a 23A current, the IGBT will be current-limited.

Current limiting is done by reducing the gate drive output voltage to maintain the 23A. This limiting occurs within 500ns of the over current and this is well within the 10µs required for the IGBT to be protected.

Third, under-voltage protection provided by IC3 prevents any gate drive if the supply is below about 8V.

Note that while IC3 is powered from 15V, the input at pin 2 can be as low as 3.3V logic level. In our circuit, a 0V to 5V signal is applied to IC3 from the PWM output of the IC1 microcontroller.

IC2a also monitors the current across the 0.01Ω shunt via a 10Ω and 470nF low-pass filter and MOSFET Q2 is used as a sample-and-hold buffer. Q2 is switched on when the PWM signal being applied to its gate is high. The MOSFET then conducts and passes the voltage that's across the 470nF capacitor through to IC2a's pin 3 input.

When the PWM signal goes low, the MOSFET is off and so the sampled voltage is stored in the 100nF capacitor.

The 15nF capacitor at the gate of Q2, in conjunction with the 1kΩ gate resistor, slows down the switch-on speed of Q2. Diode D2 switches off the MOSFET more quickly when the PWM goes low.

The slow switching of Q2 is needed to reduce voltage feed-through from the gate to the drain and source. Feed-through occurs each time the gate is switched and the sudden voltage change is capacitively coupled to the drain and source.

This effect is minimised by reducing the switch-on rate and also having a low-impedance source to the MOSFET. Low impedance is ensured using the 0.01Ω shunt, the 10Ω series resistor and 470µF capacitor.

Note that internal to Q2 is an intrinsic diode that allows conduction of current from the source to the drain. While Q2 could be connected in this circuit with the drain and source reversed, connecting this way would allow the 100nF

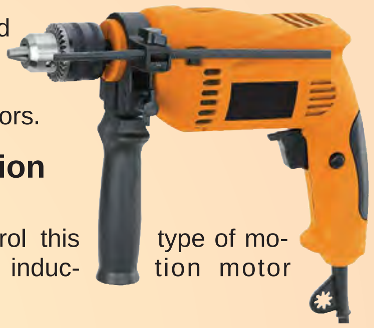
What motors can – *and cannot* – be controlled?

We've noted elsewhere in this article that this controller suits the vast majority of power tools and appliances (which use universal motors – series-wound motors with brushes). Incidentally, they're called universal motors because they can operate on both AC and DC.

But how do you make sure that your power tool or appliance is a universal motor and not an induction motor? As we also said before, induction motors **must not be used** with this speed controller. One clue is that most universal motors are quite noisy compared to induction motors. However, this is only a guide and it's certainly not foolproof.

In many power tools you can easily identify that the motor has brushes and a commutator and you see sparking from the brushes and that shows that the motor is a universal type. But if you can't see the brushes, you can also get a clue from the nameplate or the instruction booklet.

So how do you identify an induction motor? Most induction motors used in domestic appliances will be 2-pole or 4-pole and always operate at a fixed speed, which is typically 2850 rpm for a 2-pole or 1440 rpm for a 4-pole unit. The speed will be on the nameplate. Bench grinders typically use 2-pole induction motors.



Controlling induction motors

If you do need to control this type of motor then use a dedicated induction motor controller.

And a reminder:

You *cannot* control the speed of any universal motor which already has an electronic speed control built in, whether part of the trigger mechanism or with a separate speed dial.

This does not include tools such as electric drills which have a two-position mechanical speed switch. You can use our speed controller with the mechanical switch set to either fast or slow.

capacitor at pin 3 of IC2a to discharge via the diode, when the shunt resistance voltage is lower than the 100nF capacitor's voltage.

IC2a amplifies the sampled voltage by about 31. The resulting voltage is read by IC1 via its AN4 input. IC1 effectively averages the voltage at AN4 over a 10ms period so as to capture a full half-wave portion of the mains cycle for current measurement.

The averaged current measurement is multiplied by the feedback setting of VR2 (which can be regarded as optional) or VR3. This multiplication value is then used to apply PWM adjustment for maintaining motor speed.

IC1 determines if VR2 is connected at each power up. If it is not, monitoring is redirected to VR3. Initially, AN2 is configured as an output that is set low (0V). Then AN2 is reconfigured as an analogue input and the voltage level is measured.

If the level is much higher than 0V then VR2 must be connected to be able to change the level. If the level is essentially unchanged, the pin is configured as an output again, but this time the output is set high (5V). Then AN2 is set as an input and the level measured. If it remains high, then the input is open. If the input is at a lower level, then VR2 must be connected.

If VR2 is not detected, pin 1 is set as a low output and VR3 is used as the feedback input. The 2.2kΩ resistor in series is there to prevent the output being shorted during testing. The 100nF capacitor is to hold voltage during testing.

The 2.2kΩ resistor and 100nF capacitor are also included to filter out noise from associated mains wiring that could be coupled in through the potentiometer's wiper wiring. The same filtering is also included for potentiometer VR1.

Over-current

IC2b compares the voltage from detection IC2a's output (pin1) against a reference set at 3.4V by the 4.7kΩ and 10kΩ resistors connected across the 5V supply. The output (pin 7) goes high when IC2a's output is higher than 3.4V.

Output from IC2b is ignored by IC1 unless the averaged current as detected at the AN4 input exceeds 15A. IC1 then begins to reduce the duty cycle of the PWM drive until the comparator output switches low.

WARNING!

This is NOT a project for the inexperienced. Do not attempt to build it unless you are familiar with working with high voltage circuits.

Physical details

The *Motor Speed Controller* is housed in a rugged diecast aluminium case, and has separate rotary speed and adjustable feedback controls.

The controller plugs into the mains via a standard IEC mains lead, while the motorised appliance plugs into a switched mains socket on the controller's case lid.

Next month

That completes the technical description of our new *Super-Smooth Full Range Universal Motor Speed Controller*. We're sure you'll agree that this one really delivers the goods. In our next issue, we'll get into the exciting part: building it!

March medley of magical LEDs

TechnoTalk

Mark Nelson

On the menu this month is synthetic food, artificial sunlight and printing electronic devices on a 3D printer. Each of the projects involves LEDs and while you can't try them at home yet, they look entirely practical and productive. So read on, as Mark Nelson sets out his eclectic selection.

WITH EVER-MORE MOUTHS to feed, the world is desperate to find low-cost 'alternative' foods, particularly in regions with low average incomes. Furthermore, because food crops can also be used for energy production (biofuel) millions of people may face hunger as food is diverted to fuel production. So, the pressure is on to find ways of producing food artificially.

Frankenstein food

Artificial food sounds unnatural. If you watched the sci-fi film *Quatermass II*, with its evil pressure domes producing 'Frankenstein food' artificially from a witches' brew of ammonia, hydrogen, nitrogen and methane, you'll remember how obnoxious it all looked and sounded. Anyone squeamish will probably not relish the idea of eating protein obtained by fermenting the *Fusarium venenatum* microfungus fed on glucose syrup with added nitrogen, which sounds just as revolting. But it is sold in every supermarket under the name 'Quorn', a well-known meat replacement product enjoyed by vegetarians. It's also used as an anonymous additive in some bakery products. The truth is, 'artificial' foods can definitely be tasty, nutritious, and healthy.

Pond slime

A promising and nutritious candidate for artificial production is algae, which need only sunlight to grow and can thrive in salty water on barren fields. Green pond slime is one of the species of algae, which may sound unpalatable, but there are many others that we're happy to eat. Carrageenan is one of them and has been used for decades in beer, desserts, milkshakes, fruit smoothies and salad dressings. Some 80 per cent of the carrageenan we consume comes from the Philippines, where it flourishes naturally, but that's a lot of food-miles for consumers in the West. Now a process for growing edible algae artificially anywhere in the world under LED 'sunlight' has been proposed by Prof Thomas Brück of the Technical University in Munich, Germany.

'Algae grow much faster than soya beans or corn. They require neither fertile ground nor pesticides and have a ten-fold higher yield per hectare and year,' he says. 'And since they are so undemanding and thrive even in salt water basins set up on barren fields, they could help solve the problems posed by using food crops for energy production.'

LEDs to the rescue

There's always a snag, however. 'Nobody can really predict whether algae from the tropics will be as productive under German light conditions as in their native environment,' says Brück. 'To begin, nobody knows whether candidates that work here would be equally successful in the bright conditions of the Sahara. But now we can test all of these things in our laboratory.'

LEDs can compensate for the low-luminance skies of Europe, but reproducing sunlight accurately in the laboratory is a challenge. This is why the university has teamed up with the Berlin-based LED manufacturer FutureLED to develop a methodology for simulating all kinds of light situations. Ultra-efficient LEDs provide light with wavelengths between 400 and 800nm and a radiation intensity of 1,000W/m^2 with an intensity distribution that models natural sunlight very closely. The various LED types can be controlled individually, allowing the researchers to program specific light spectra.

Biofuel spin-off too

Closer investigation of specific types of algae led the scientists to discover a variety of promising products. Many algae produce intermediate chemicals and synthesise fats that could be converted into fuels. But even within a single species, the ability to produce specific products varies widely. 'In our investigations we keep seeing huge differences in productivity,' concludes Brück. 'So we have to identify not only the right species, but must also cultivate the candidates with the highest productivity.'

The scheme is part of the larger 'AlgenFlugKraft' (algae-powered flight) project, which is also investigating catalytic conversion of biomass, algae processing, fat separation and hydrogen/biomass production. This looks so promising that the German state of Bavaria and the Airbus Group are funding the project to the tune of 12 million euro (£9.5 million) and a dedicated technical research centre is already under construction south of Munich.

3D optics

No, not another way of watching movies at the cinema, but a remarkable development in 3D printing and electronics. Researchers at Princeton University (New Jersey, US) have managed to embed tiny light-emitting diodes into a standard contact lens, allowing the device to project beams of coloured light. You're probably wondering what's the use of this and you'd be right to ask, because assistant professor Michael McAlpine, the lead researcher on this project, concedes that the lens is not designed for actual use, but to successfully demonstrate the ability to 3D print electronics into complex shapes and materials.

He explains: 'This shows that we can use 3D printing to create complex electronics, including semiconductors. We were able to print an entire device, in this case an LED, as well as the hard contact lens. To create the LEDs that generate the coloured light, his team used quantum dots (also known as nanoparticles) as an ink.'

Complementary technologies

Clever as 3D printing is, McAlpine says it is unlikely to replace traditional manufacturing in electronics any time soon; instead, they are complementary technologies with very different strengths. Traditional manufacturing, which uses lithography to create electronic components, is a fast and efficient way to make multiple copies with a very high reliability. Manufacturers are using 3D printing, which is slow but easy to change and customise, to create moulds and patterns for rapid prototyping. 'Trying to print a cell phone is probably not the way to go,' he said. 'It is customisation that gives the power to 3D printing.' In this case, the researchers were able to custom 3D print electronics on a contact lens by first scanning the lens, and feeding the geometric information back into the printer, enabling the conformal 3D printing of an LED on the contact lens.

Stereo Echo & Reverb Unit

By NICHOLAS VINEN

Based on the *Stereo Audio Delay* featured in the February 2015 issue, this modified unit can be used to provide adjustable echo or reverberation for recording or public address (PA) work. By using revised software and slight changes to the circuitry, we show how the same hardware can provide these different functions. We'll also describe some extra features that can be useful in either mode.

IN ESSENCE, the *Stereo Audio Delay* described in the February 2015 issue consists of an analogue-to-digital converter (ADC) and a digital-to-analogue converter (DAC), with a PIC32 microcontroller processing the digital audio stream between the two. This microcontroller has a large internal RAM (128KB) which, together with an optional external 1MB SRAM chip, can be used for buffering and manipulating the audio data stream.

By controlling how much of this memory is used for buffering, the PIC32 can delay the audio by a variable amount. But it can also process the audio data – perform some sort of filtering, for example.

In fact, providing an adjustable echo effect requires only a small amount of additional processing compared to what's needed for audio delay; we simply need to mix a proportion of the delayed audio back into the input signal.

This simulates a real (acoustic) echo, whereby sound waves travel a significant distance, resulting in a time delay (since sound travels at around 340m/s at sea level). The attenuation of expanding sound waves travelling through a significant volume of air, along with the losses inherent in reflections off less-than-perfect surfaces, result in the volume of the echo being lower than that of the original sound.

The echo itself has an echo, so that a single transient sound has a number of echoes, spaced equally apart in time and with a decaying sound level. This aspect of echo is also simulated by the above simple method. That's because by mixing an attenuated version of the delayed signal back into the input signal, that echo itself is delayed and attenuated, and so on *ad infinitum* until the volume has decayed so far that it is no longer audible (see Fig.1).

Fig.3 shows the circuit of the *Stereo Echo & Reverb Unit*. It's basically just the *Stereo Audio Delay* described in February 2015 with various optional extra bits added on (plus the revised software for the micro). Provision was made on the original PCB to accept these extra bits, so you don't have to start from scratch with a new board. Instead, it's just a matter of building the PCB as originally described and adding the extra parts.

Enabling echo mode

As it stands, the February 2015 unit can be switched from delay mode

The new unit is built on the same PCB as the *Stereo Audio Delay Unit* described in February 2015. It's just a matter of adding a few extra parts and using revised software.

Features and specifications

- Adjustable stereo echo or reverb with interval of 0-640ms
- Echo delay and attenuation adjustable via front panel knobs
- Optional defeat switch connection for foot pedal; can be configured as normally on or normally off
- THD+N: <0.03% (typically <0.02%), 20Hz-20kHz (20Hz-22kHz bandwidth)
- Signal-to-noise ratio: typically >76dB (line inputs/outputs)
- Optimal line input signal range: 0.5-2V RMS
- Line output signal: 1V RMS
- Input impedance: 4-6kΩ (line input), 8.2kΩ (microphone input)
- Power supply: 7.5-12V DC or 3.8-6.5V DC, depending on configuration; current drain 60-80mA
- Microphone input: 20-50mV input for full scale output, signal-to-noise ratio 67dB.
- Headphone output: drives 8-32Ω at up to 50mW, THD+N 0.05% @ 10mW/32Ω, volume adjustable in 1dB steps

to echo mode by placing a shorting jumper across pins 3 and 4 of CON7, the ICSP header.

When the unit is powered up, the software briefly attempts to pull pin 4 high and checks its state. With no jumper plugged into CON7, this pin will be sensed as high and so the unit will perform its default task, which is to provide audio delay. However, if pins 3 and 4 are shorted, pin 4 will remain low despite the pull-up and so echo mode is activated.

As explained last month, pins 4 and 5 of CON7 are normally PWM signals generated by the microcontroller, which can be measured in order to determine the configured delay in milliseconds. But if the software detects that pin 4 is shorted to ground at start-up, it disables this PWM output in order to avoid driving this short circuit. You can measure the echo delay at pin 5; in echo mode, the unit will only operate in stereo so there is only one delay to measure.

The other reason that echo will only operate in stereo is that in this mode, VR1 (or VR3) is used to set the echo delay while VR2 (or VR4), if present, sets the echo attenuation. If neither VR2 nor VR4 are installed, then the attenuation is set to 12dB.

As with the regular delay mode, a delay of up to 600ms is available without the external SRAM chip IC3 fitted or up to six seconds with IC3 in place. But 600ms is quite a long delay and should be sufficient for most echo effects.

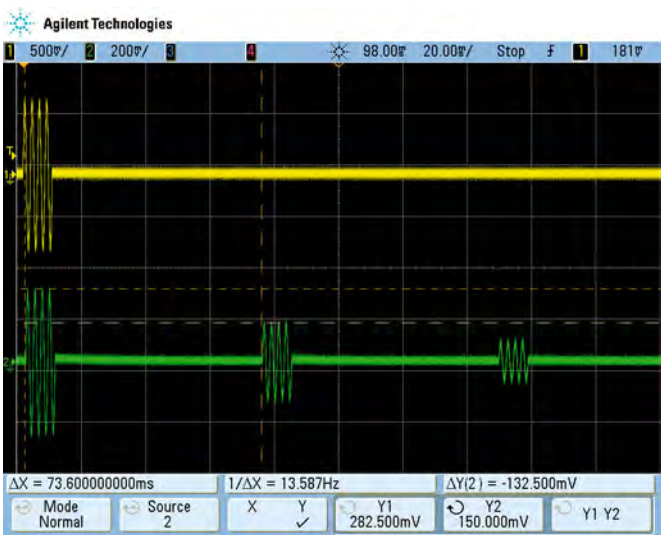


Fig.1: this scope grab shows the input (yellow) and output (green) signals when the unit is set to echo mode with a delay of approximately 70ms and an attenuation of around 6dB. The initial burst is output immediately at a somewhat reduced level, followed by echoes, of which the first two are shown. Each is lower in amplitude compared to the previous echo.

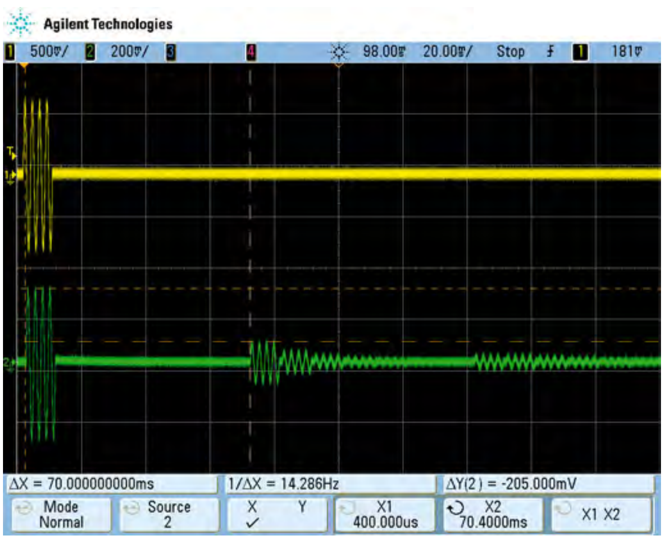
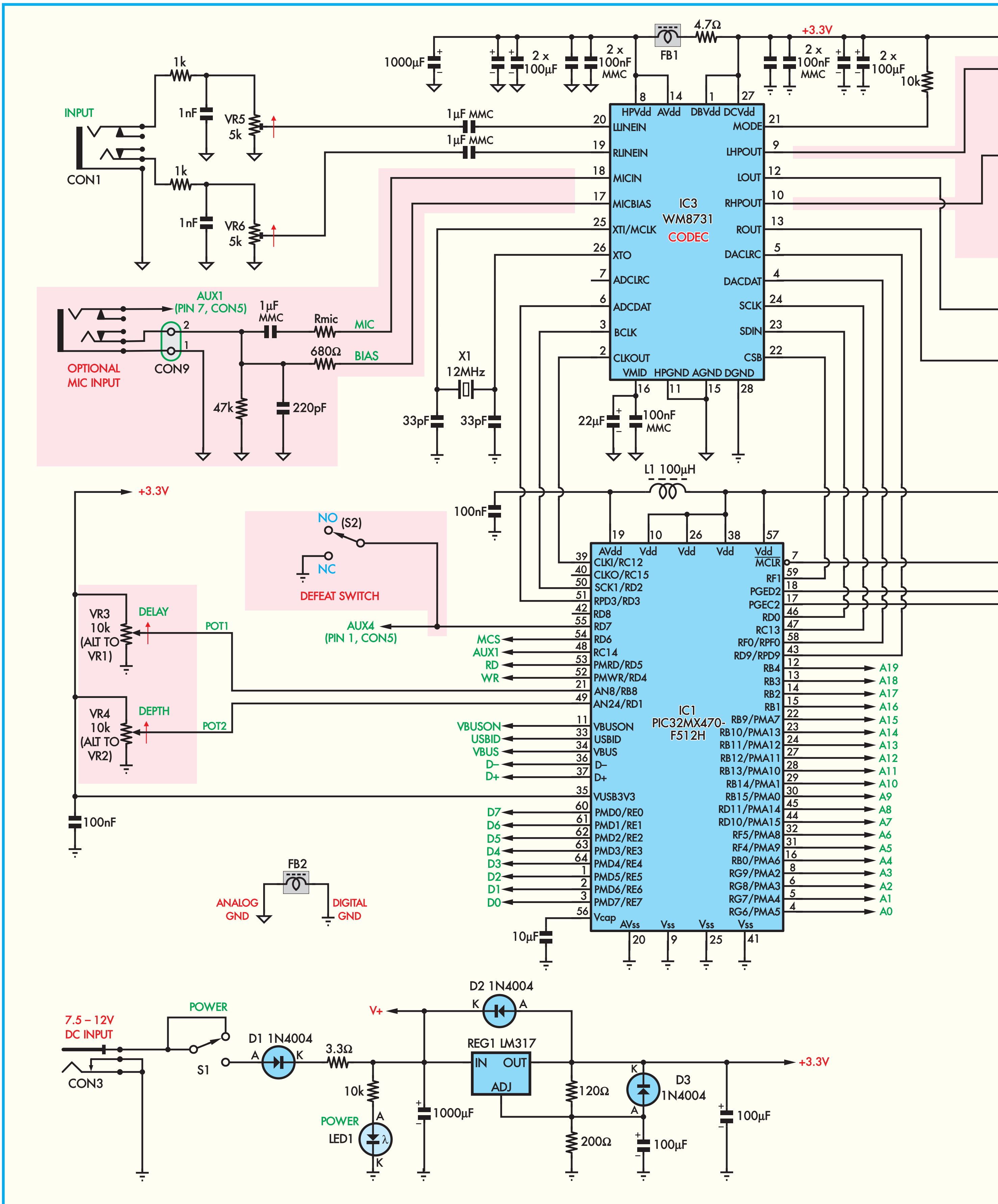


Fig.2: this scope grab shows the same waveforms as in Fig.1, but this time with reverb mode enabled and using a similar delay. In this case, the echoes are even lower in amplitude but they are followed almost immediately by a further series of 'sub-echoes' which themselves decay fairly rapidly. This makes for a more complex echo effect with greater 'depth'.



STEREO ECHO/REVERBERATION UNIT

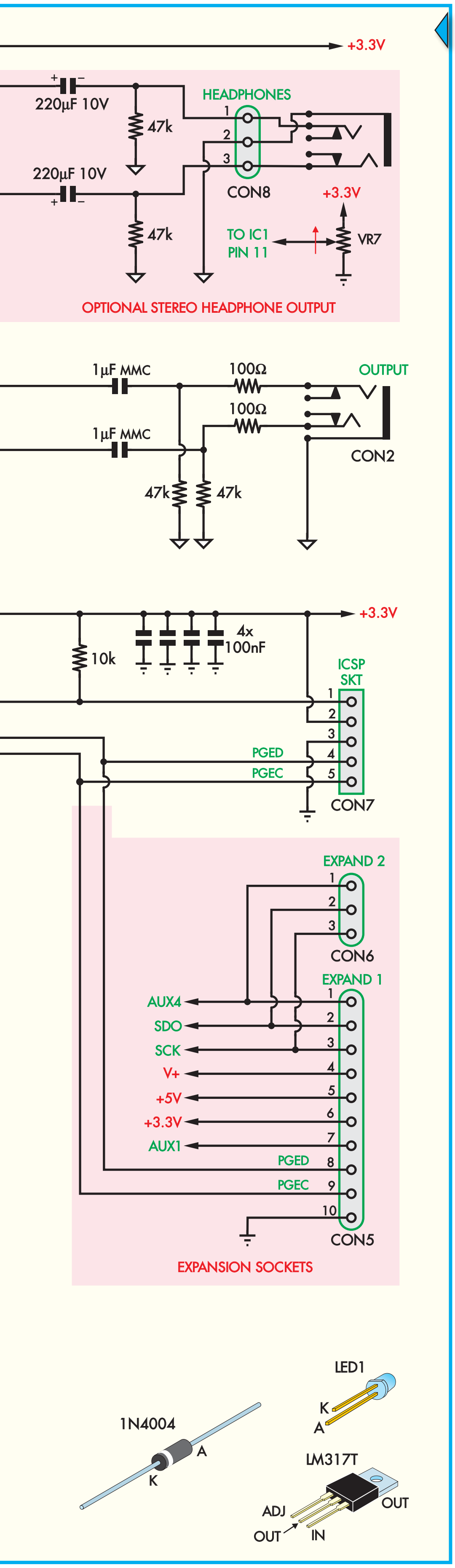


Fig.3: the *Stereo Echo & Reverb Unit* circuit – it's basically the same as the *Stereo Audio Delay Unit* published in February 2015, but with the circuitry highlighted with a red background added (plus revised software for IC1).

So really it's just a matter of building the unit as described in the February 2015 issue, with VR3 and VR4 fitted, installing the jumper on pins 3 and 4 of CON7, powering it up and then adjusting the knobs until you get an echo effect that you are happy with.

We have produced new front and rear panel labels (Fig.5) with positions marked to drill the extra holes for VR3 and VR4. Positions are also marked for a headphone volume control and output socket, which we'll explain later. These panels can either be copied or downloaded from the *EPE* website.

Defeat switch

If you are building this as an echo effects unit for musical performances, you will need a way to switch it on and off. To shut off the echo, we simply pull input RD7 (pin 55) of IC1 low; it is internally biased high by a weak current source. This pin is labelled as AUX4 on the circuit diagram (Fig.3) and is wired to a pad on the PCB at the top, near the middle (see Fig.4).

These pads are designed to suit an SPDT right-angle pushbutton switch, but for musical performances, having a button on the unit isn't very practical. Instead, we suggest fitting a 3.5mm phono jack socket to the rear panel of the unit and wiring it to the two switch connections on the PCB via a 3-way header (ie, middle-pin unused). A foot switch can then be plugged in via a length of cable fitted with a 3.5mm jack plug.

Foot switches generally have a double-throw switch, with three terminals: COM, NC (normally closed) and NO (normally open). If you wire the plug tip and sleeve to the COM and NC terminals, pushing on the foot switch will enable the echo effect and it will stop when you lift your foot off. This is the most logical way to wire it. However, you could also wire the plug to the COM and NO terminals and then the echo effect will be disabled by pressing on the switch and re-enabled by lifting off.

The wiring arrangement for the foot switch socket is shown in Fig.4, along with some extra wiring we'll describe later. This is also shown on the circuit diagram (Fig.3).

Our revised rear panel artwork includes a hole position marked for the foot switch socket and an associated label. Note that the position shown has been chosen to avoid interference between the panel-mounted socket and trimpots VR5 and VR6.

Reverb

Echo is basically a simple form of reverb (or reverberation). In a space such as a concert hall, there won't be just a single echo duration for sounds originating on the stage and being heard by people sitting in the audience. Instead, there will be many different paths that the sound can take. The direct path is the shortest and gives the least attenuation, but sounds also bounce off various surfaces before reaching the listener and each path will have its own delay and (probably frequency-dependent) attenuation.

Professional reverberation units can provide many different options, to simulate halls of various different configurations. In this unit, we've stuck with a simple approach which gives a more complex (and audibly distinct) response than a simple echo without being terribly processor intensive or having a lot of parameters to tweak.

Essentially, to obtain the reverb effect, we add an echo with a short time delay to the sound, then take the resulting sound and process that with a much longer delay. This simulates a large space where there are multiple paths for the sound to bounce around, each with a slightly different length, and thus the echoes arrive at slightly different times.

To enable reverb mode, pins 3 and 5 of CON7 must be shorted. Pin 4 should be left open and can be used to measure the set delay.

Note that since pins 3 and 5 aren't adjacent, you can't use a shorting block to do this. The trick is to use a 3-pin female header and solder a short length of wire between the two outside pins and then plug this into CON7. Of course, you could solder a wire directly to pins 3 and 5 of CON7 but then it's harder to disconnect.

As with echo, reverb mode only operates in stereo and the adjustments are identical. The long delay is adjusted as

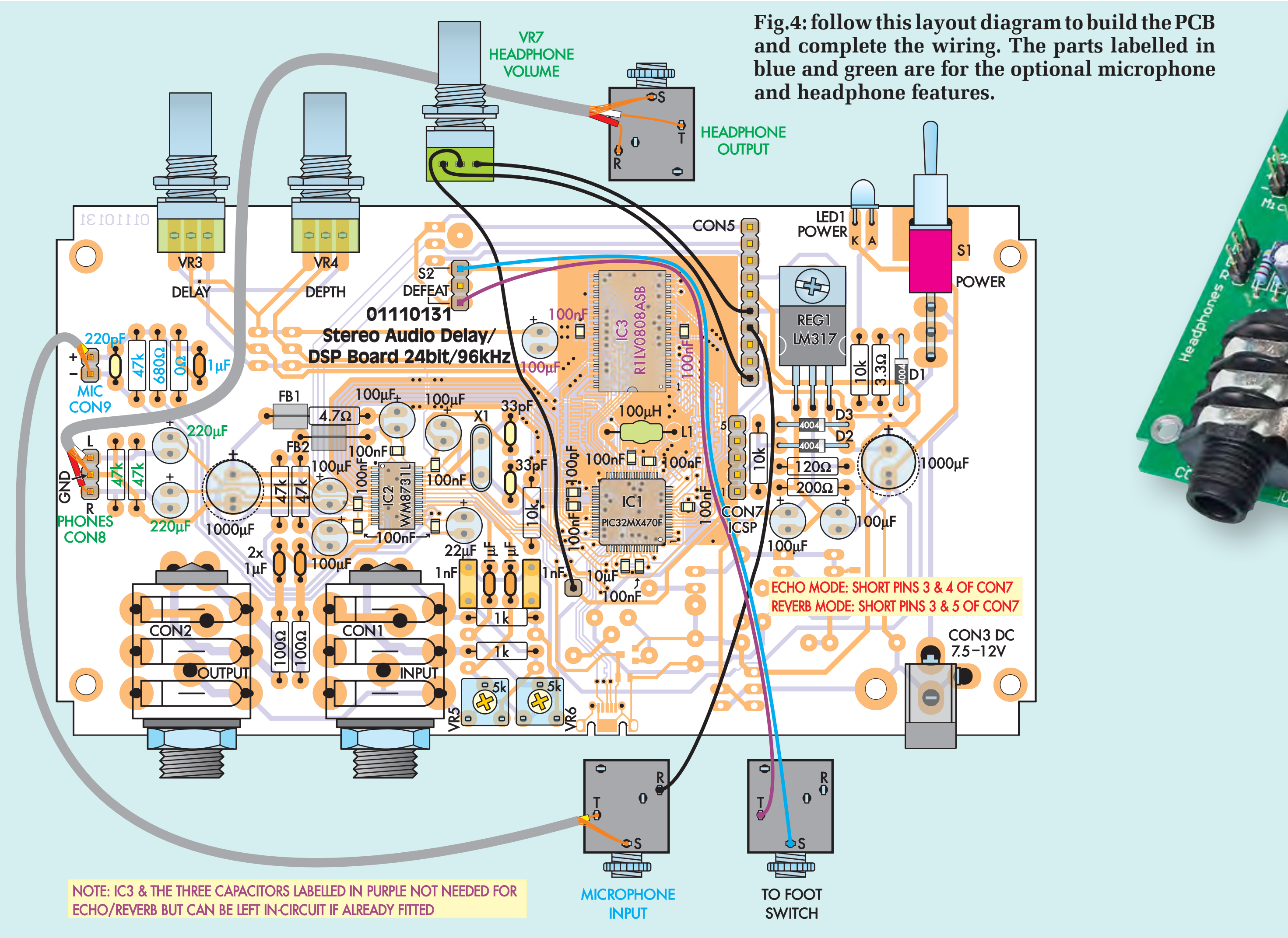


Fig.4: follow this layout diagram to build the PCB and complete the wiring. The parts labelled in blue and green are for the optional microphone and headphone features.

for echo mode while the short delay is automatically set to be 1/8th as long. So if you select a 200ms long delay, the short delay will be 25ms.

The same attenuation setting is used for both short and long delays and as with echo, this can be adjusted with VR3/VR4; otherwise it's set to 12dB by default.

Echo/reverb switching

An SPDT switch can be used if you want to be able to switch between echo and reverb modes. To do this, first connect its common terminal to pin 3 of CON7 (or another ground connection point) via a series 1kΩ resistor. The two remaining switch terminals then go to pins 4 and 5 of CON7.

The 1kΩ resistor is necessary to prevent a dead short to the PWM output if switching is done while the unit is on. Note that since the unit only checks the state of these pins at power-up, you would then have to switch the unit off and then on again to complete the changeover.

Headphone and mic support

In the February 2015 issue, the following pins of CODEC IC2 were unconnected: LHPOUT, RHPOUT, MICIN and MICBIAS. These pins can be used for a microphone input and/or a stereo headphone output – see Fig.3.

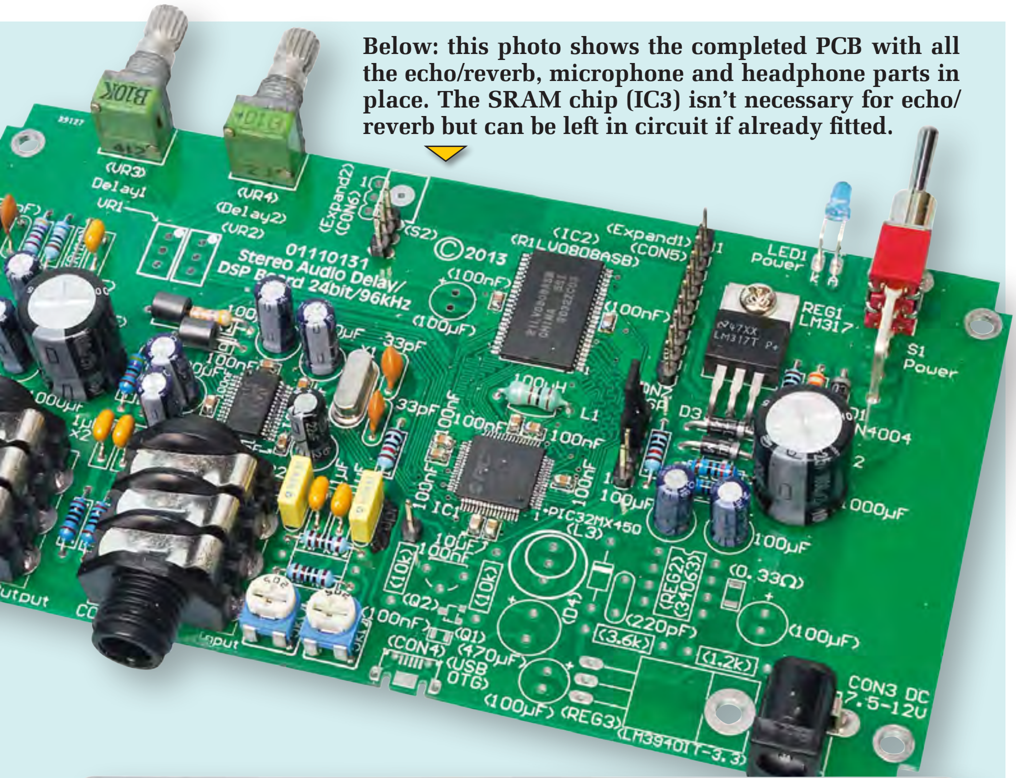
This allows you to take advantage of the headphone amplifier with digital volume control and the microphone amplifier with electret bias built into the IC. When a microphone is connected, the microcontroller detects this and automatically switches from sampling the line inputs to sampling the microphone input.

By the way, the microphone amplifier in IC2 is quite a bit noisier than a good external microphone amplifier (which could be connected to the line inputs) but you may find it suitable for some uses (see the spec. panel).

To add a headphone output, it's just a matter of installing the additional parts labelled in green on the PCB layout (Fig.4). This consists of two 220µF DC-blocking electrolytic capacitors and their associated 47kΩ resistors, plus 3-pin header CON8. A panel-mounted 3.5mm phono socket is then wired back to this header. We've shown a 2-core shielded cable for this connection, but it doesn't really need to be shielded.

You also need to wire up an extra potentiometer (VR7) to allow the volume to be controlled. This volume pot is also mounted on the front panel, next to the other two pots, and wired to 10-way pin header CON5 (which must also be installed) and to a single pin soldered to a pad just below and to the left of IC1.

By wiring the pot this way, we're connecting in virtually the same manner as VR3 and VR4 – ie, across the 3.3V supply rail. The wiper is connected



Below: this photo shows the completed PCB with all the echo/reverb, microphone and headphone parts in place. The SRAM chip (IC3) isn't necessary for echo/reverb but can be left in circuit if already fitted.

Above: the fully-assembled PCB mounted inside the case and with the wiring completed. Use cable ties to secure the wiring as shown.

to pin 11 of IC1 via the single pin connection shown, which is the only remaining free ADC-capable input of IC1. This connection is also indicated on the circuit diagram (Fig.3).

The revised software (0111013B.hex) for microcontroller IC1 automatically detects when this pot is present and if it is, constantly samples the voltage at pin 11. If this voltage changes, IC1 sends a command to CODEC IC2 to adjust the headphone output volume.

Microphone input

The extra circuitry required to hook up a microphone is also quite simple. As shown on Fig.3, the signal from the microphone is fed in via a 1µF non-polarised capacitor. The resistor labelled 'Rmic' is normally 0Ω, which sets the microphone gain to 26dB. However, if this is too much gain, you can reduce it somewhat by using a non-zero-value resistor.

A 39kΩ resistor for Rmic will reduce the gain to 20dB, while a 15kΩ resistor will give a gain of approximately 23dB. For maximum gain, if you don't have a 0Ω resistor, use a wire link instead.

If using an unpowered electret microphone, it will require a small bias current to operate. In this case, the 680Ω resistor should be fitted and the bias current will come from IC2's MICBIAS output which is enabled by default when a microphone is plugged in. The 47kΩ resistor to ground provides 0V DC bias for the microphone when there is no bias current, while a 220pF capacitor provides a small amount of RF filtering.

The micro detects when a microphone is plugged in by monitoring input pin RC14 (pin 48) which is connected to a track labelled 'AUX1' via pin 7 of 10-way pin header CON5. As explained previously, this header was intended at the time for future

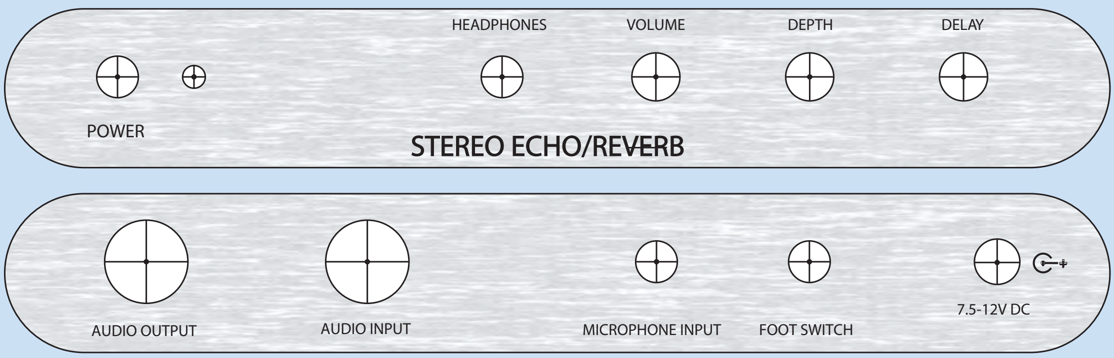


Fig.5: these two artworks can be copied and used as drilling templates for the front and rear panels. They can also be downloaded as a PDF file from the *EPE* website.

The additional 3.5mm microphone input and foot-switch jack sockets are mounted on the rear panel. A 7.5-12V DC plugpack powers the unit.

Parts List

1 double-sided PCB available from the *EPE PCB Service*, coded 01110131, 148 × 80mm
1 ABS plastic instrument case, 155 × 86 × 30mm
1 set front and rear panel labels
4 No.4 × 6mm self-tapping screws
1 12MHz HC-49 crystal (X1)
1 100µH axial RF inductor (L1)
1 10kΩ multi-turn vertical trimpot (VR1) OR 1 × 10kΩ 9mm horizontal potentiometer (VR3)
2 5kΩ horizontal mini trimpots (VR5,VR6)
2 6.35mm PCB-mount stereo switched jack sockets (CON1,CON2)
1 5-way pin header, 2.54mm pitch (CON7)
1 PCB-mount SPDT right-angle toggle switch (S1)
1 DC plugpack, 7.5-12V, 100mA+
2 4mm ferrite suppression beads
1 PCB-mount switched DC socket to suit plugpack
1 M3 × 6mm machine screw/ nut

Semiconductors
1 PIC32MX470F512H-I/PT 32-bit microcontroller programmed with 0111013B.hex (IC1)
1 WM8731SEDS 24-bit 96kHz stereo CODEC (IC2) (Element14 1776264)
1 LM317T adjustable regulator (REG1)
1 3mm blue LED (LED1)
3 1N4004 diodes (D1-D3)

Capacitors
2 1000µF 25V electrolytic
6 100µF 16V electrolytic
1 22µF 16V electrolytic
1 10µF 6.3V 0805 SMD ceramic
4 1µF 50V monolithic ceramic
11 100nF 6.3V 0805 SMD ceramic
2 1nF MKT
2 33pF ceramic disc

Resistors (0.25W, 1%)

2 47kΩ	1 120Ω
3 10kΩ	2 100Ω
2 1kΩ	1 4.7Ω 0.5W 5%
1 200Ω	1 3.3Ω 0.5W 5%

Add-on Features

For echo/reverb:
1 3.5mm panel-mount stereo jack socket
1 3-way pin header
1 jumper shunt
1 100mm length 2-strand ribbon cable
1 2-core cable with 3.5mm jack plug at one end (length as required)
1 foot switch

For headphone output:
1 3.5mm panel-mount stereo jack socket
1 10kΩ 9mm panel-mount linear potentiometer
1 small knob to suit
2 220µF 10V electrolytic capacitors
2 47kΩ 0.25W resistors
1 100mm length 2-core shielded cable or 3-strand ribbon cable
1 100mm length 3-strand ribbon cable
1 14-way (or more) snappable pin header

For microphone input:
1 3.5mm panel-mount stereo jack socket
1 1µF multi-layer ceramic capacitor
1 220pF ceramic capacitor
1 47kΩ 0.25W resistor
1 680Ω 0.25W resistor
1 2-way pin header
1 100mm length shielded cable
1 100mm length ribbon cable strand or light-duty hookup wire

For low-voltage supply:
1 LM3940IT-3.3 or TS2940-3.3 low-dropout 3.3V regulator (REG2)
1 1N5819 1A Schottky diode (D1)
1 470µF 10V electrolytic capacitor
1 100µF 16V electrolytic capacitor
Delete REG1 and associated parts

Note: microcontroller IC1 must be programmed with revised software (ie, 0111013B.hex) for echo/reverb and the other add-on features to work.

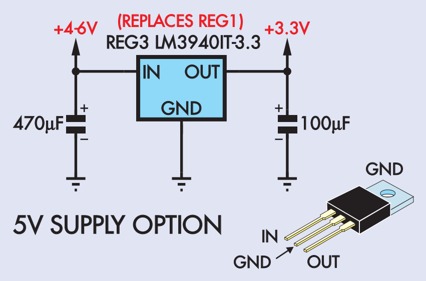


Fig.6: the unit can be powered from a 4-6V DC supply by replacing REG1 with an LM3940IT-3.3 low-dropout regulator as shown here.

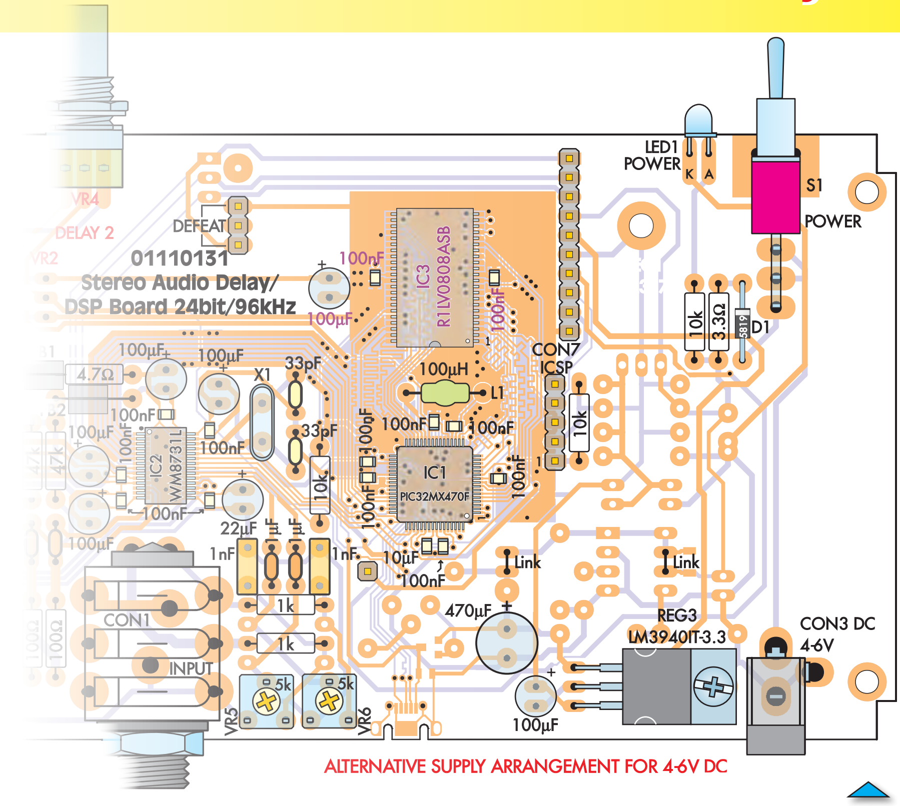


Fig.7: follow this PCB parts layout diagram to install the parts for the 4-6V power supply option. Note that D1 must be changed to a 1N5819 Schottky type.

expansion. The micro enables a weak internal pull-up on this pin, which is connected to the ring terminal of the microphone input.

Since the microphone input is mono, when a mono plug is inserted, this will short the ring and sleeve terminals. The sleeve is connected to ground and so AUX1 is pulled low. The micro mutes the input for half a second when this input changes state. If, after this period, the input is low then the microphone input is selected. Otherwise, the line input is used.

Thus, if a microphone is plugged in, the unit automatically switches to that as the signal source and when it is removed, it automatically switches back to the line inputs. Because the microphone input is mono, the same signal is sent to both audio output channels.

5V/battery operation

The power supply for the unit is based on a series polarity protection diode (D1) and an LM317 regulator (REG1) configured to provide a 3.3V output. This requires an input voltage of 6V or more (preferably 7.5-12V) to ensure proper output regulation.

However, as stated in February 2015, it's possible to reconfigure the unit to run from 3.8-6.5V. This makes it suitable for use with USB power (4.25-5.5V), a single Li-ion or Li-Po cell, or four standard cells (alkaline or rechargeable).

This alternative power supply arrangement is shown in Fig.6 and the parts layout diagram of Fig.7. Basically, an LM3940IT-3.3 fixed low-dropout regulator (REG3) is used instead of the LM317T, along with a couple of wire links to get power to it. In addition, the 1N4004 reverse polarity protection diode (D1) is replaced with a 1N5819 Schottky diode – the latter has a much lower forward voltage drop.

This gives a minimum operating voltage of around 3.6V, so if you power the unit from a Li-ion or Li-Po cell, the cell will be pretty much fully discharged before the circuit ceases normal operation (in practice, it will probably operate down to at least 3.3V but without supply regulation).

Note that while this arrangement allows the unit to run off lower input voltages, damage may occur if more than 7V is applied, even briefly. So if using a plugpack with this new arrangement, **be sure to measure its actual unloaded output voltage rather than relying on its nominal rating. A 5V unregulated plugpack could easily put out more than 7V at light load.**

Building it

The *Stereo Echo & Reverb Unit* is built on the same PCB as the *Dual-Channel Audio Delay* (February 2015), which is available from the *EPE PCB Service*.

Note that the some of the new features, such as the microphone input option and headphone output, can also be used for the delay function.

In fact, once the unit is completed, it can be easily switched between providing a delay and operating in echo/reverb mode. So the same hardware can fulfill either role.

Either way, you will need to program the micro with the revised firmware (ie, 0111013B.hex) which can be downloaded from the *EPE* website. If you buy a pre-programmed chip, it will come with this version.

Basically, it's just a matter of first building the PCB as described in the February 2015 article. You then simply add the extra parts to the PCB for the microphone and/or headphone options and complete the wiring as shown in Fig.4.

Don't forget to link the appropriate pins on CON7 to enable echo or reverb mode. Link pins 3 and 4 for echo mode (use a shorting jumper), or pins 3 and 5 for reverb mode (use a 3-way female header with the outside terminals linked).

If you want to run the unit from a 4-6V supply, then build the power supply section as shown above in Fig.7. Don't forget that diode D1 (near switch S1) must be a 1N5819 Schottky type.

Fig.5 shows the revised front and rear panels and these can be used as drilling templates for the extra holes required for the potentiometers and the stereo jack sockets. Note the headphone volume pot (VR7) is mounted directly on the front panel rather than the PCB.

Completing the "Tiny Tim" Stereo Amplifier

Part 3

Nicholas Vinen

In this final instalment we finish building the "*Tiny Tim" Stereo Amplifier* by fitting all the modules into the case and wiring it up. We'll also look at testing the unit, its final performance and some other useful tidbits.

At this stage, you will have finished building the main amplifier PCB and power supply and you should also have prepared the case, including drilling holes in the base for mounting the modules. But before we screw them in, it's easiest to do some of the wiring first.

Start with the wiring between the two chassis-mount RCA sockets, the slide switch, the RCA plugs for the DAC and the leads that connect to the amplifier PCB. This wiring is shown in the upper-left corner of Fig.6 on page 24 of the February 2015 issue.

Strip and tin the wires to go to the PCB but leave these loose; the rest of the wiring can be completed in place.

Note that depending on how close you have mounted the RCA sockets to the slide switch, it may be impractical to use shielded cable for these connections, in which case you will have to use ordinary hook-up wire instead. In this case, keep the wires as short as possible and run the two signal wires close to the ground wire(s) to minimise hum pick-up.

Fitting the DAC

With that done you can then mount the DAC board. As explained last month, to save space we fitted ours directly above the RCA sockets and slide switch and we used a combination of various nylon tapped spacers, nuts and screws to support it.

Essentially, what you need to do is fit the DAC connectors and switch through the rear panel holes you made earlier and measure how high the DAC sits above the bottom of the case, then pick the next shortest tapped M3 spacers you can get.

Experiment with how many M3 nuts or washers you need to fit to the screws before attaching the spacers so that the DAC board naturally rests on these spacers when it is in place.

It's then just a matter of using a few more nylon M3 screws to hold it in place on the top. The holes on the DAC board are a bit bigger than usual for M3 screws, but the screw heads should be sufficiently large to hold it down. Otherwise, use nylon washers under the screw heads. You can then plug the two RCA cables you soldered earlier into the DAC outputs.

Output and pot wiring

The next step is to fit the front panel components and connect wires in preparation for the final assembly. This wiring consists of the following runs; again, refer to Fig.6 in the February 2015 article:

1) Two red wires from the left and right channel pins on the headphone socket, long enough to reach the amplifier board, plus a black ground wire of a similar length.
2) Two long red wires from the switched left and right channel pins on the headphone socket to run along the bottom of the case and back to the two red binding posts. Remember to slip a couple of pieces

of heatshrink tubing over each wire before soldering them to the binding posts, and it's also a good idea to wrap the exposed copper strands securely around the binding post pin before soldering it (which will require a hot iron).

That done, slide the heatshrink over the solder joint and shrink it down, then repeat for a double insulating layer (see photos). We attached several adhesive plastic wire clips to the bottom of the case to hold these wires in place, roughly along the paths shown in Fig.6.

3) Two black wires from the black binding posts, long enough to reach the rear of the amplifier board and connect to the ground plane. These should have two layers of heatshrink insulation over the solder joints.
4) Two stereo shielded wires soldered to the volume control pot, long enough to reach to the pot connections on the amplifier board. Wire these as per Fig.6 last month.

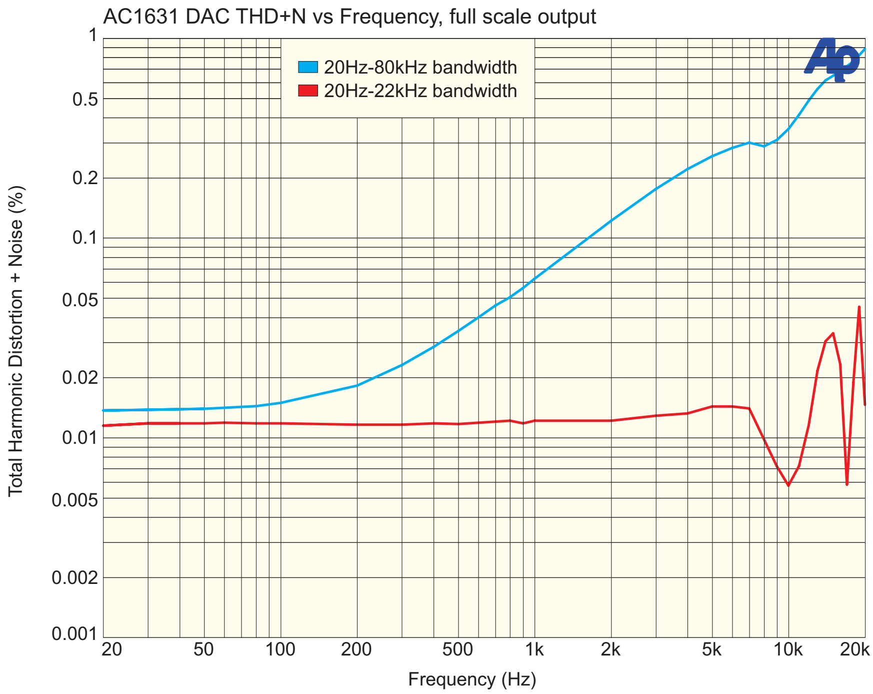
Mounting modules and testing

First, fit the power supply module in place by screwing its four tapped spacers into the bottom of the case. Use three short steel M3 machine screws and a nylon M3 machine screw for the right-rear corner, ie, the mounting post which already has a nylon screw in the top.

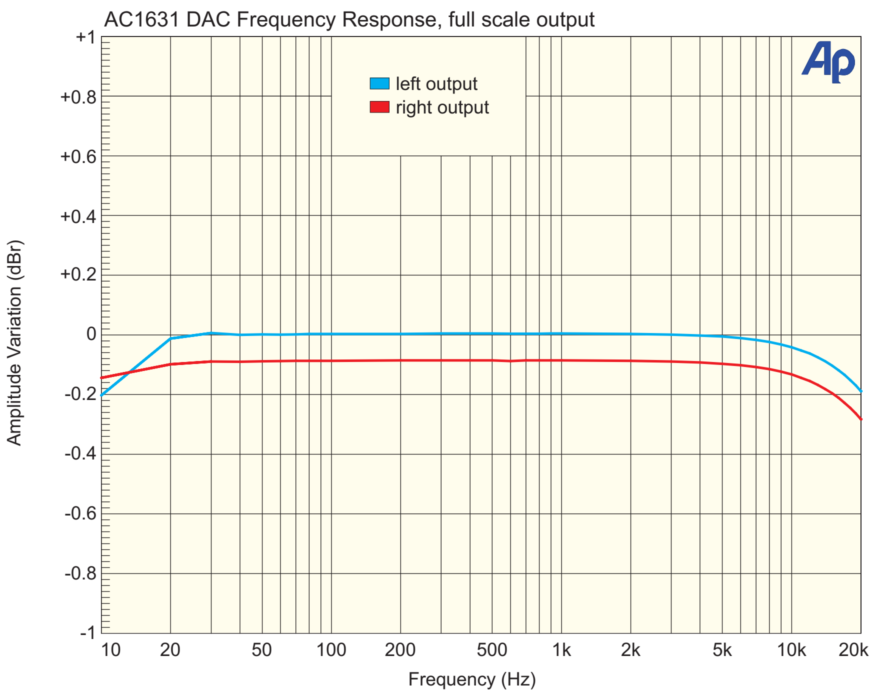
Cut a 60 × 40mm piece of fibre insulation (eg, Presspahn) and then score and fold it 45mm from one end. Drill two holes in this to correspond to the two holes in the bottom of the case, near the power supply board and attach it using M3 nylon machine screws and nuts, as shown in the photos. This prevents any wires which may come loose from contacting any of the mains-potential components on the PCB.

Connect the mains cable to the left-most pin header terminal and feed it through its grommet at the rear of the case (ie, the one that it went through originally).

At this point, with the power supply in the case, it's probably a good idea to check that it is working properly – so, plug the switch in and check that it is properly isolated. To do this, set your DMM to a high ohms range (ie, megohms) and connect one probe to the mains plug live pin and one to an exposed piece of metal on the chassis. Check that there is no connection (it should read 'oL' or similar).



Distortion vs frequency at full-scale output for the AC1631 DAC module. It's an oversampling type DAC, rather than delta-sigma, hence the rather steep rise in distortion with increasing frequency. But when the output is filtered with the 20Hz-22kHz bandpass filter (red trace, simulating human ear response), most of the distortion harmonics are eliminated. Comparing this graph to the others shows that when using a digital input, the DAC is generally the limiting factor in performance.



The frequency response of the AC1631 DAC is pretty flat, being down by only 0.2dB at the high end (20kHz) and virtually flat to 20Hz at the low end. Note that it does not handle Dolby Digital, DTS or other compressed audio streams. So, if connected to a TV set or disc player, the unit should be configured to output a linear PCM stereo digital signal. Most disc players and many TV sets offer a 'down-mixing' option, specifically to allow the digital audio output to be connected to devices like this.

Reproduced from last month, this photo shows the placement and interconnection of the PCBs within the *Tiny Tim Stereo Amplifier*.

Repeat the same test with the neutral pin. Then check, with the power switch on, that there is no connection from the mains live pin to any of the three terminal block outputs on the power supply PCB. This verifies that the transformer insulation is intact. Assuming that's all OK, switch the DMM to DC voltage measurement mode and check that the power supply fuse cover and adjacent Presspahn shield are in place, plug in the mains cord and turn it on.

Without touching the mains section of the power supply board, measure between the middle pin of the terminal block and either side. You should get readings of approximately ±20V (likely a bit higher). Switch off and check that these drop to near 0V within about 30 seconds. This confirms that the power supply board is working and you can then switch off and unplug the mains and then the mains switch from the power supply board.

Note that with some terminal blocks, there may not be a good connection to the screw on top when there is no wire inserted, so it's best to probe the wire openings if possible.

Amplifier module installation

Before fitting the amplifier module to the case, make sure you have soldered the three power supply wires as shown in Fig.6 last month, and that they are long enough to reach the power supply output terminals when it is in the case. A 2-wire cable should also be attached for the 12V DC output, as described last month. If you fitted sockets to the amplifier board, plug in the ICs now, but make sure their pin 1 dot lines up with the notch on the socket.

You can now mount the amplifier module using four tapped spacers and eight short M3 machine screws. The *MiniReg* board is mounted in a similar manner (note that no heatsinking or regulator tab connection is required) and the two-pin header you wired to the amplifier board's 12V rail earlier can now be plugged into the *MiniReg's* input. Check the polarity, ie, ensure the grounds of the two boards are continuous, eg, from the OUTPUT – pin of CON4 on the *Minireg* to the tinplate shield on the amplifier board.

You can also plug in the power LED into the *Minireg* now. But we don't want to connect the power supply directly to the amplifier PCB just yet, with the exception of the 0V (black) wire which can go to the central output on the power supply board. Leave the other two (red and blue) loose for now.

Next solder the remaining wires to the PC pins on the amplifier board, specifically the six from the pot, three from the headphone socket, four for the inputs (from the chassis-mount slide switch) and two from the black binding posts.

It's a good idea to slip a short length of heatshrink tubing over each wire before soldering (slide it far enough along the wire so that it doesn't shrink from the heat) and then shrink it down over the solder joint when it's cooled to provide some strain relief.

More testing

We now want to check whether the amplifier module is working and the best way to do this is to temporarily connect a couple of 100Ω 5W safety resistors in series with the supply leads so that if something is wrong, you will have time to switch power off before any damage occurs. This also reduces the chance of a problem when adjusting the amplifier's quiescent current.

If you have enough room, you can insert one lead of each safety resistor into one of the terminal block outputs on the power supply board, screw it down and bend it up so that the resistors stick up vertically. It's then just a matter of running a clip lead from the other end of each resistor to the appropriate power supply wire for the amplifier module.

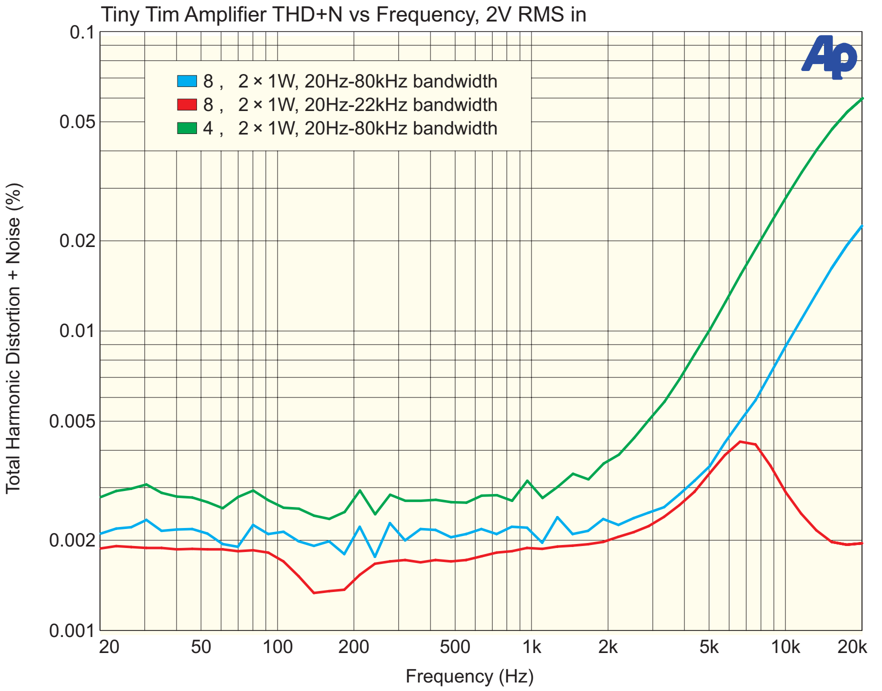
Make sure that the clip lead from the red wire goes to the safety resistor at the positive output terminal on the power supply, which is furthest from the corner of the board. If necessary, use clip leads at both ends of the safety resistors and they can sit outside the case. But regardless, make sure that the exposed metal of the alligator clips can not make contact with anything else – a good way to ensure this is to temporarily wrap them in electrical tape.

For now, do not connect the DC output from the *MiniReg* board to the DAC's power supply input socket. Re-connect the mains power switch, do a final check to make sure there are no stray wires that could short to anything (especially near the power supply board!) and turn trimpots VR2 and VR3 on the amplifier board fully anti-clockwise.

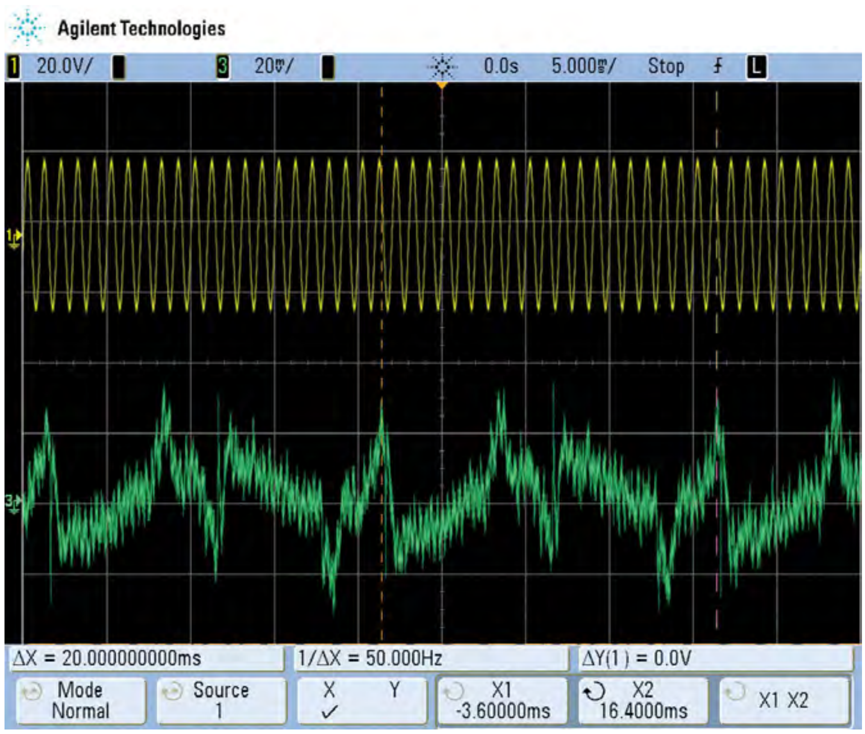
You can now plug the unit back into the mains, switch it on and check the voltage across each safety resistor using a DMM set to DC volts mode.

Don't go near the mains side of the power supply. You should get a reading below 10V in each case (typically around 8-9V); if not, switch off immediately and check for faults in the wiring.

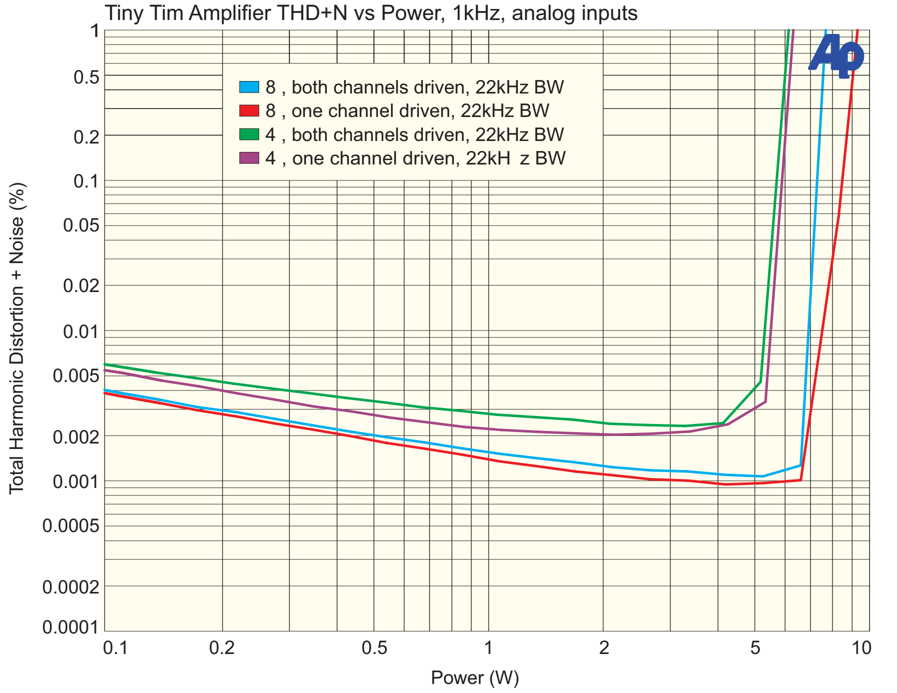
If the wiring looks OK but the voltages are too high, there there is probably a problem with the component installation or the modifications to the amplifier board. Assuming the voltages are OK, measure the voltage between each pair of red/black binding posts (ie, the output offset voltage). It should be below 20mV. If it's much higher than that, there is



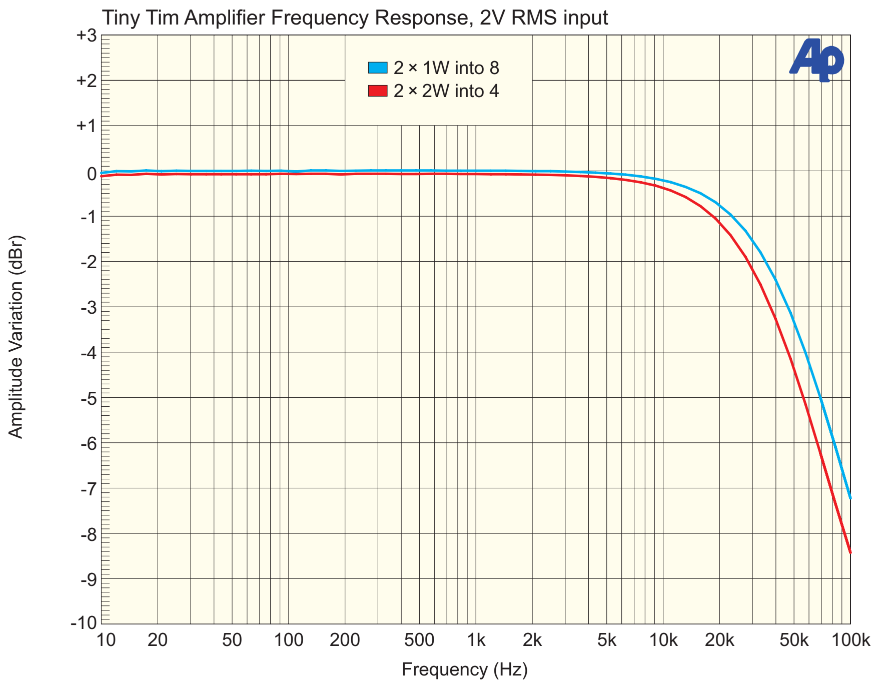
Distortion versus frequency from the completed amplifier under a variety of conditions. This is somewhat higher than what was shown for the amplifier module/power supply combination in the January 2015 issue. This is almost entirely due to increased hum and rectifier buzz pick-up now that the boards are mounted close together in the case. If we measure the distortion with a 400Hz high-pass filter to eliminate mains 50Hz hum and its immediate harmonics, the readings drop substantially, to around 0.0006%.



The distortion residual of the amplifier output at 1W with both channels driven into an 8Ω load (green) compared to the output itself (yellow). As you can see, it is mainly a combination of 50Hz, 100Hz and even-order harmonics of these frequencies, indicating that it's due to hum pick-up from the power supply. The actual distortion products at 2kHz and above can be seen superimposed on this waveform at a much lower level.



Distortion versus power for a 1kHz signal under various conditions. As is typical, distortion is lower into 8Ω loads than 4Ω due to the lower output current for the same power level. Continuous power output is below 10W, but music power (ie, the power available for short bursts) is higher than this, at about 10W for both 4Ω and 8Ω speakers with both channels driven. Note that despite the level of hum measured, even with the volume turned up and our ear very close to the speaker we could barely make it out (inputs must be terminated for this test).



Frequency response of the complete amplifier. Note that this shows a much greater roll-off at the high end (down by about 1dB at 20kHz) compared to the graph published in the January 2015 issue. That's because since taking the earlier measurements, we decided to increase the input filter capacitors to 4.7nF to give better attenuation for the harmonics in the DAC output. You could lower this value slightly to give a slightly flatter high-frequency response, but then it would be less effective at attenuating DAC switching noise.

a fault, so switch off and check your work carefully.

Otherwise, now is also a good time to check the output of the *MiniReg* board, either at CON4 or if you have plugged it in, the DC plug (with the red probe inserted through the end and the black in contact with the outside of the barrel).

Turn its adjustment trimpot and check that the output voltage varies. You can then set it to 6V±0.1V.

Next, connect the DMM between TP1 and TP2 on the amplifier board and slowly rotate VR2 clockwise. The voltage should start out low (just a few millivolts) and rise as you turn the pot. Stop once it reaches 15mV.

Note that we indicated a reading of 30mV in the main circuit diagram in the January 2015 issue, but we have found that the heatsinks run a bit hot at idle; 20mV is plenty of bias in practice. We're setting it to 15mV now because it will increase a bit once the safety resistors have been removed. Repeat this procedure for TP3/TP4 and trimpot VR3.

Check the voltage across the safety resistors again. It should have increased to around 12V and they will be getting a little warm.

Having passed those tests, the amplifier board is likely working – but if you want to be really sure, you can do a live signal test by connecting a pair of speakers and some sort of signal source. But if you do this with the lid open, you need to be careful not to go anywhere near the power supply. In fact, we would switch off and unplug the unit while connecting the speakers and signal source.

Of course, with this sort of test it's always a good idea to turn the volume control right down before switching back on and advance it slowly. While the power switch is off you should also check that the input selector slide switch is in the appropriate position for the analogue inputs.

With the safety resistors in place, only a small amount of power will be available, but you should be able to get clean audio at a reasonable volume.

You can then switch off, unplug the mains cord and wire the amplifier module directly to the power supply, making sure you hook up the wires to the same terminals as you used earlier. You can now also connect the output of the *MiniReg* to the DAC board.

That should complete the wiring. To keep it neat and safe, tie all the cables

into bundles or to adjacent posts so that they can't move and break loose should the unit be subject to vibration or shock.

If in doubt, refer to our photos (including those published last month) to see how we did it. Your completed unit should look much like ours, although obviously it will vary somewhat depending on which case you used.

Now is a good time to repeat the live signal test but this time without the safety resistors, you should have the full power output of up to 10W per channel available. Once it's warmed up a little bit, re-adjust VR2 and VR3 to get 20mV across the associated test posts.

Assuming it all works and sounds good, you can switch off, unplug the mains cord and attach the lid, volume knob and any other ancillaries to complete the unit, such as feet. Make sure the mains cord is properly anchored using the original method once the lid is in place; in some cases the lid helps to hold the cordgrip grommet in place.

The accompanying graphs show the performance of the completed unit and the integrated DAC. These measurements include power supply noise, hum, RF pick-up in the wiring and so on, so they aren't quite as good as the performance of the amplifier module itself, but still pretty good and we think you will find the sound quality is 'up to scratch'.

Depending on what speakers you are using, you may want to consider adding a bass extender to your new Hi-Fi setup.

Modifying the DAC for more output

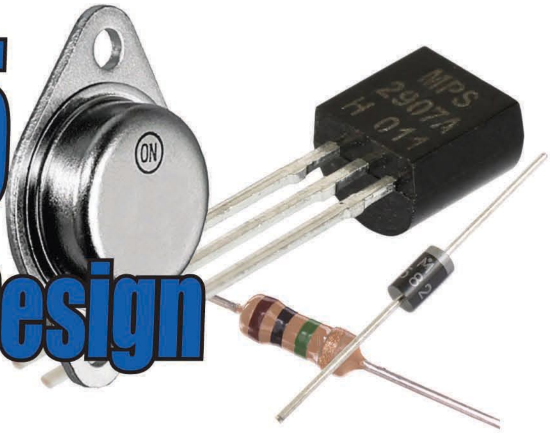
The pre-built DAC board we have used in this project (Jaycar AC1631) has an output of around 1V RMS, while most CD/DVD/Blu-Ray players and high-end DACs have an output closer to 2V RMS. This is generally not a big problem, but it does mean that if you are switching between the analogue and digital inputs, you will need to adjust the volume control each time.

Reader Gavin Krautz wrote to us to explain that he has this DAC and grew tired of constantly changing volume levels when switching inputs; he came up with a simple way to increase the DAC output level to around 2V RMS. As he explains below:

'The DAC contains a BH3544 headphone amplifier to drive the outputs, which has a default gain of 6dB. However, its gain can be reduced by inserting resistors in series with the signal going to pins 3 and 5 of the IC. In the Jaycar DAC, these resistors (R25 and R27) are 90kΩ, which sets it gain to 0dB (ie, unity). The formula given for the gain is 6dB + 20.log10(90kΩ÷(90kΩ + R_{ip})).

'This means you can increase the output gain by up to 6dB by changing these resistors. I initially considered making the output gain switchable, or shunting R25 and R27 to increase the gain, but in the end I simply shorted them out to restore the 6dB default and I have been very happy with the result.'

Teach-In 2015



Discrete Linear Circuit Design

Part 2 : Testing and measurements

by Mike and Richard Tooley

Welcome to *Teach-In 2015* – this series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery. Each part of our *Teach-In* series is devoted to a different aspect of discrete linear circuit design, such as modelling and simulation, measurement and testing, noise and distortion. In last month's introductory instalment, *Discover* introduced discrete semiconductors while *Knowledge Base* described the theory relating to the amplifiers in which bipolar junction transistors (BJT) are used. Our practical feature, *Get Real*, described the design and construction of a simple pre-amplifier and our *Special Feature* provided you with an introduction to SPICE and how to get started with the powerful and versatile TINA Design Suite.

Introduction

In this month's *Teach-In 2015 Knowledge Base*, we explain hybrid parameters and show you how they can be used to predict the performance of a transistor amplifier. *Discover* will introduce you to the use of virtual test instruments – digital storage 'scopes in particular. Our practical feature, *Get Real*, will explain how we used powerful SPICE-based software to design and develop the pre-amplifier module that we described last month. We will also describe a variety of basic tests and measurements that can be carried out on the 'real' pre-amplifier circuit using low-cost virtual test instruments. All you will need to get started is a pre-amplifier to test, a 9V PP3 battery, some test leads and an ordinary PC fitted with a sound card.

Discover: Virtual test instruments

In recent years, an increasingly powerful yet affordable range of virtual instruments has become available. Unlike 'real' test instruments, such as hand-held multimeters, signal generators and bench oscilloscopes, virtual instruments make use of a computer (usually a laptop or desktop machine) for signal processing and display. The screen of the computer is usually arranged so that it provides the same set of controls and types of display that you might find in a real test instrument, but with the added advantage of the ability to save and recall settings and also to capture data and save it for analysis at some later time.

When using a computer as the basis of a virtual test instrument there are two basic approaches. One involves the use of external hardware and the other is based on the sound card that's already built into a PC (or which can easily be added if none is available). We will briefly describe each of these potential solutions in relation to the requirements for a digital storage oscilloscope (DSO).

Computers with external hardware

The first and more powerful method of realising a DSO is that of using external hardware (effectively a high-speed multi-channel ADC) connected to a PC via a USB (or other) port. The necessary software is usually supplied on CD-ROM or can be downloaded from the manufacturer's website. It is important to note that, although the external hardware cannot usually be used without the appropriate software, some manufacturers also make available software drivers that will allow you to control the oscilloscope and capture data for use in your own applications. However, for most of us this isn't an option since the supplied software will usually outperform anything that we can write ourselves.

Several types of external hardware-based digital storage scopes (DSO) are currently available. These can be conveniently arranged into three different basic categories according to their application:

- Low-cost DSO
- High-speed DSO
- High-resolution DSO

Unfortunately, there is often some confusion between the last two classes of DSO, so to put the record straight, a high-speed DSO is primarily designed for examining waveforms that are rapidly changing. Such an instrument does not necessarily provide high-resolution measurement. On the other hand, a high-resolution DSO is useful for displaying waveforms with a high degree of precision, but it may not be suitable for examining fast-changing waveforms. The difference between these two types of DSO should become a little clearer later on, but first it is worth explaining how the sampling rate of a DSO impacts on the frequency range over which it can be reliably used.

Sampling rate

The upper signal frequency limit of a DSO is determined primarily by the rate at which it can sample an incoming signal. Typical sampling rates for different types of DSO are shown in Table 2.1.

In order to display waveforms with reasonable accuracy it is normally suggested that the sampling rate should be at least twice and preferably up to five times the highest signal frequency. Thus, in order to display a 10MHz signal with any degree of accuracy a sampling rate of 50M samples per second is recommended.

The 'five times rule' merits a little more explanation. When sampling signals in a

Table 2.1 Typical sampling rates for various classes of DSO

Type of DSO	Typical sampling rate	Typical applications
Low-cost (entry level)	20K to 200K per second	Power and audio frequency
High-speed	100M to 1000M per second	RF, video and pulse testing
High-resolution	20M to 200M per second	High accuracy measurement

Table 2.2 Various types of signal and their typical bandwidth requirements

Signal	Bandwidth required (approx.)
Low-frequency and power	DC to 10kHz
Audio frequency (general)	20Hz to 20kHz
Audio frequency (speech)	300Hz to 3.4kHz
Audio frequency (high-quality)	10Hz to 50kHz
Square and pulse waveforms (up to 10kHz)	DC to 100kHz
Fast pulses with small rise-times	DC to 1MHz
Video	DC to 10MHz
Radio (LF, MF and HF)	DC to 50MHz

digital-to-analogue converter, we usually apply the Nyquist criterion. This states that the sampling frequency must be at least *twice* the highest analogue signal frequency. Unfortunately, this no longer applies in the case of a DSO where we need to sample at an even faster rate in order to accurately display the signal. In practice, we would need a minimum of about five points within a single cycle of a sampled waveform in order to reproduce it with reasonable fidelity. The sampling rate should therefore be at least *five* times that of highest signal frequency.

A special case exists with dual channel DSOs. Here the sampling rate may be shared between the two channels. Thus, an effective sampling rate of 20M samples per second might equate to 10M samples per second for *each* of the two channels. In such a case the upper frequency limit would not be 4MHz, but only 2MHz.

The approximate bandwidth required to display different types of signals with reasonable precision is given in Table 2.2.

When determining bandwidth requirements, the authors' rule of thumb is that for sinusoidal signals the bandwidth should be at least double that of the highest signal frequency, while for square wave and pulse signals, the bandwidth should be around ten times that of the highest signal frequency component. It is also worth remembering that most manufacturers define the bandwidth of an instrument as the frequency at which a sine wave input signal will fall to 0.707 of its true amplitude (ie, the –3 dB point). To put this into context, at the cut-off frequency, any displayed trace at the –3 dB point will be in error by a whopping 29%!

Resolution

Resolution is usually expressed in terms of the number of bits used in the conversion process; the more bits used in the conversion process the greater the number of discrete voltage levels that can be resolved. The relationship is as follows:

$$x = 2^n$$

where x is the number of discrete voltage levels and n is the number of bits. Thus, each time we use one additional bit in the conversion process we double the resolution of the DSO, as shown in Table 2.3.

Table 2.3 Relationship between resolution and signal accuracy

Number of bits, n	Number of discrete voltage levels, x
8	256
10	1,024
12	4,096
16	65,536

Buffer memory

A DSO stores its captured waveform samples in a buffer memory. Hence, for a given sampling rate, the size of this memory buffer will determine for how long the DSO can capture a signal before its buffer memory becomes full. The relationship between sampling rate and buffer memory capacity is important. A DSO with a high sampling rate but small memory will only be able to use its full sampling rate on the highest time-base ranges.

To put this into context it's worth considering a simple example. Assume that we need to display 10,000 cycles of a 10MHz square wave. This signal will occur in a time frame of 1ms. If we applying the five-times rule we would need a bandwidth of at least 50MHz to display this signal accurately.

To reconstruct the square wave we would need a minimum of about five samples per cycle, so a minimum sampling rate would be 5 × 10MHz = 50M samples per second. To capture data at the rate of 50M samples per second for a time interval of 1ms would require a memory that can store 50,000 samples. If each sample uses 16-bits we would need 100KB of extremely fast memory!

Accuracy

The measurement accuracy of a DSO is specified in terms of the smallest voltage change that can be measured and, by implication, it will depends on the actual measuring range selected. For example, on the 1V range an 8-bit DSO is able to detect a voltage change of one 1/256V (or about 4mV). For most measurement applications this will prove to be perfectly adequate, as it amounts to an accuracy of about 0.4% of full-scale.

Low-cost DSO are primarily designed for low frequency signals (typically signals up to around 20kHz) and are usually able to sample their signals at rates of between 10K and 100K samples per second. Resolution is usually limited to either 8-bits or 12-bits (corresponding to 256 and 4,096 discrete voltage levels respectively). Such instruments may be either single or dual channel and, because of their low cost and simplicity, they are ideal for educational purposes as well as making basic measurements. Most low-cost DSOs provide all of the functionality associated with a conventional scope, but with many additional software-driven features at a fraction of the price of a comparable conventional test instrument. That said, there are two important limitations with an inexpensive DSO that you need to be aware of:

1. Bandwidth is generally limited, so they are only suitable for examining low-frequency and audio signals
2. Resolution is generally limited to 8-bits or 256 discrete steps of voltage

High-speed DSOs are usually dual-channel instruments that are designed to replace conventional general purpose oscilloscopes (but with the added advantage that captured data can be stored for subsequent processing and analysis). These instruments have all the features associated with a conventional scope, including trigger selection, time-base and voltage ranges, and an ability to operate in X-Y mode.

Features offered by computer-based oscilloscopes

Additional features available with a computer-based instrument that are not usually available from conventional bench-top scopes include the ability to capture transient signals and save waveforms for future analysis. The ability to analyse a signal in terms of its frequency spectrum is yet another feature that is only possible with a DSO.

The technique of Fast Fourier Transformation (FFT) of the data captured by a DSO makes it well suited to produce frequency spectrum displays. Such displays can be used to investigate the harmonic content of waveforms, as well as the relationship between several signals within a composite waveform. This type of display is just not possible with a conventional analogue oscilloscope, where only voltage/time displays are available.

Auto-ranging is another very useful feature that is often provided with a DSO. If you regularly use a conventional scope for a variety of measurements you will know only too well how many times you need to make adjustments to the vertical sensitivity of the instrument. To have this adjustment performed for you automatically is an absolute boon!

Resolution and bandwidth

High-resolution DSOs are used for precision applications where it is necessary to faithfully reproduce a waveform and also to be able to perform an accurate analysis of noise floor and harmonic content. Typical applications include small-signal work and high-quality audio.

Unlike the low-cost DSO, which typically has 8-bit resolution and poor DC accuracy, high resolution DSOs are usually accurate to better than 1% and have either 12-bit or 16-bit resolution. This makes them ideal for audio, noise and vibration measurements. The increased

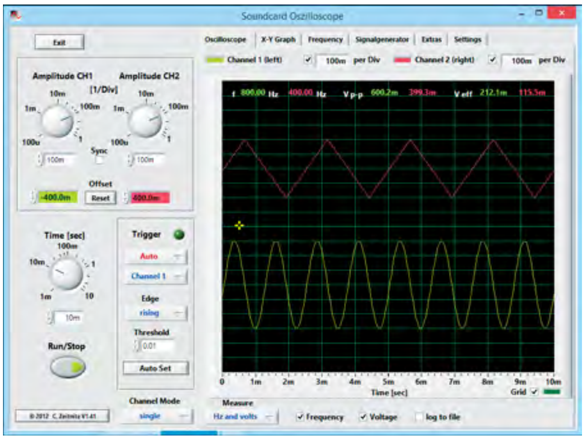


Fig.2.1 Soundcard Scope from Christian Zeitnitz is available for free download provided that it is only used for private (non-commercial) and public education. This excellent software will work with any 16-bit sound card at a sampling rate of 44.1kHz. It also has an integrated signal generator that produces sine, triangle, square, sawtooth and white noise output waveforms

resolution also allows the instrument to be used as a spectrum analyser with very wide dynamic range (up to 100dB). This feature is ideal for performing noise and distortion measurements on low-level analogue circuits and high-fidelity equipment generally (such as CD and MP3 players).

Bandwidth alone is not enough to ensure that a DSO can accurately capture a high frequency signal. The goal of manufacturers is to achieve a flat frequency response. This response is sometimes referred to as a Maximally Flat Envelope Delay (MFED). A frequency response of this type delivers excellent pulse fidelity with minimal overshoot, undershoot and ringing.

It is important to remember that, if the input signal is not a pure sine wave it will comprise a number of higher frequency harmonics. For example, a square wave will contain odd harmonics that have levels that become progressively reduced as their frequency increases. Thus, to display a 1kHz square wave accurately you need to take into account the fact that there will be signal components present at 3kHz, 5kHz, 7kHz, 9kHz, 11kHz, and so on.

As mentioned earlier, it is wise to purchase a DSO with a bandwidth that is five times higher than the maximum frequency of any signal that you wish to measure. Note, however, that with some instruments the specified bandwidth is not available on all voltage ranges, so it is worth checking the manufacturer's specification carefully.

Most DSOs have two different sampling rates (or modes) depending on the signal being measured: real-time and equivalent-time repetitive sampling (often referred to as ETS). Note, however that since ETS works by building up the waveform from a series of successive acquisitions, it will only provide you with a true picture of what's going on if the signal that you are measuring is both stable *and* repetitive.

Sound card oscilloscopes

The alternative solution to the use of external hardware is that of using a well-specified PC sound card. This can be extremely cost-effective because the sound card will provide the necessary ADC function. That said, it will only provide you with an accurate measurement capability that can cope with signals over a relatively low range of frequency (for example, 100Hz to 4kHz). Any PC fitted with a standard sound card (or equivalent sound card chipset integrated with the motherboard) can be used to record and playback analogue signals and this facility will also allow it to be used as the basis of a simple low-specification DSO. Furthermore, because the sound card will have a DAC (for playback) as well as an ADC (for recoding) it can also be used as a simple programmable signal source.

Most modern sound cards can usually be configured for stereo or mono operation, with either 16 or 24-bits and sampling rates of 44.1kHz (CD quality) up to 192kHz. Note that more expensive plug-in (eg PCI cards) and external (professional quality) sound systems can usually achieve higher sample rates than those provided by integrated sound cards. The stereo capability makes it possible to have two independent channels (as with most conventional scopes). In addition, the output capability of a PC sound card can be used to provide test signals within the audio and low-frequency spectrum. Sound card oscilloscope software usually provides this function in addition to waveform display of signals applied to the stereo input channels.

Limitations of sound card-based oscilloscopes

As mentioned earlier, the upper signal frequency limit of a DSO is determined primarily by the rate at which it can sample an incoming signal. Using a maximum sound card sampling rate of 44.1kHz a sinusoidal signal at 20kHz can be displayed with acceptable accuracy (the Nyquist criterion that we met earlier). However, in order to display non-sinusoidal signals faithfully, we need to sample at a much higher rate if we are to accurately display the signal. In practice, we would need a minimum of about five points within a single cycle of a sampled waveform in order to reproduce it with approximate fidelity. With a fixed 44.1kHz sampling rate this would suggest that audio frequency signals of up to 4kHz can be reasonably faithfully displayed using a sound card. It is also necessary to take into account the fact that a sound card interface is AC coupled and therefore is unable to respond to DC levels.

Provided that you are willing to accept the bandwidth limitation (typically 10Hz to no more than about 4kHz with a 44.1kHz sampling rate) the ability to convert an analogue signal into digital stored data makes it possible to use a sound card as the ADC component of a very basic DSO. With the aid of appropriate software (available either for free download or at minimal cost) the stored data produced by a sound card can be made to display on a screen in much the same way as it appears on the screen of a conventional DSO. Furthermore, the use of Fast Fourier Transformation (FFT) software makes it possible to produce a frequency-domain display as well as the more conventional time-domain display.

While it is possible to use a sound card without any external interface (most

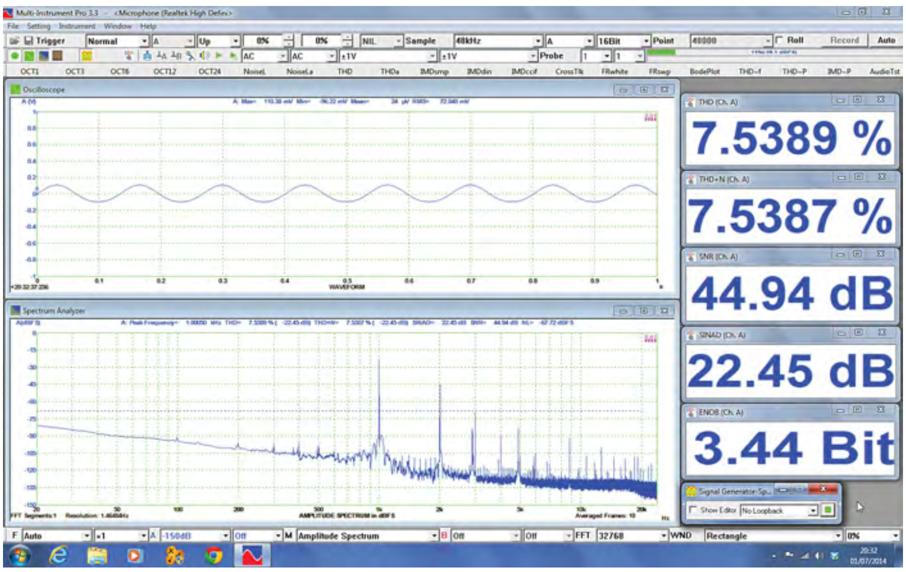


Fig.2.2 Sound card oscilloscopes can carry out a wide variety of measurements including harmonic distortion and signal-to-noise ratio. This Multi-instrument display from the Virtins sound card oscilloscope shows simultaneous displays of voltage against time (top) and voltage against frequency (bottom)

Fig.2.3 The Virtins sound card oscilloscope probes are designed to be connected directly to the audio input jack connectors of a PC. For accurate measurements, the probes can be calibrated using the Virtins software

sound cards have ample sensitivity, so additional gain from a pre-amplifier stage is unlikely to be required) there are several good reasons for using purpose-designed probes (see Fig.2.3) or some other form of input interface:

- The usual input impedance of a sound card (typically around 50kΩ) is appropriate for most audio equipment, but is too low for many measurement applications
- The usual input impedance of a sound card is too low for use with conventional scope probes, which are usually designed to work with standard 1MΩ oscilloscope inputs
- The sensitivity of most sound cards varies according to software gain settings, and the size/resolution of PC displays also tends to vary widely, hence some method of calibration is essential if accurate measurements are to be made.

Provided that you are willing to work within the above limitations you will find that a sound card-based virtual oscilloscope can be a very cost-effective solution for displaying and analysing signals in linear discrete circuits. Later in this month's *Teach-In* we will show you just how this is done.

Knowledge Base: Hybrid parameters

In last month's *Teach-In 2015* we introduced transistor amplifiers. This month we take a closer look at an equivalent circuit that we can use to predict the performance of a BJT amplifier. We will start by looking at the characteristic curves that show how a transistor responds to different values of voltage and current applied to its terminals.

BJT characteristics

The characteristics of a transistor are often presented in the form of a set of graphs that show the relationship that exists between the voltages and currents present at a transistor's terminals. For transistors designed primarily for use in linear applications, these graphs are invariably included in manufacturers' data sheets and they usually include:

- Input characteristic (base current plotted against base-emitter voltage with the collector-emitter voltage held constant)
- Output characteristic (collector current plotted against collector-emitter voltage with the base current held constant)
- Transfer characteristic (collector current plotted against base current with collector-emitter voltage held constant).

A typical input characteristic (I_B plotted against V_{BE}) for a small-signal general-purpose NPN transistor operating in common-emitter mode is shown in Fig.2.4(a). This characteristic shows that very little base current flows until the base-emitter voltage (V_{BE}) exceeds 0.6V. Thereafter, the base current increases rapidly. Note that this characteristic bears a close resemblance to the forward part of the characteristic for a silicon diode, which is hardly surprising when you recall that the base-emitter junction is forward biased, see Fig.1.2 last month.

A typical set of output characteristics (I_C plotted against V_{CE}) for a small-signal general-purpose NPN transistor operating in common-emitter mode is shown in Fig.2.4(b). Note that this characteristic comprises a family of curves, each relating to a different value of base current (I_B). It is worth taking a little time to get familiar with this characteristic as we shall be putting it to good use later on. In particular it is important to note the 'knee' that occurs at values of V_{CE} of about 2V. Also, note how the curves become flattened above this value with the collector current (I_C) not changing very greatly for a comparatively large change in collector-emitter voltage (V_{CE}).

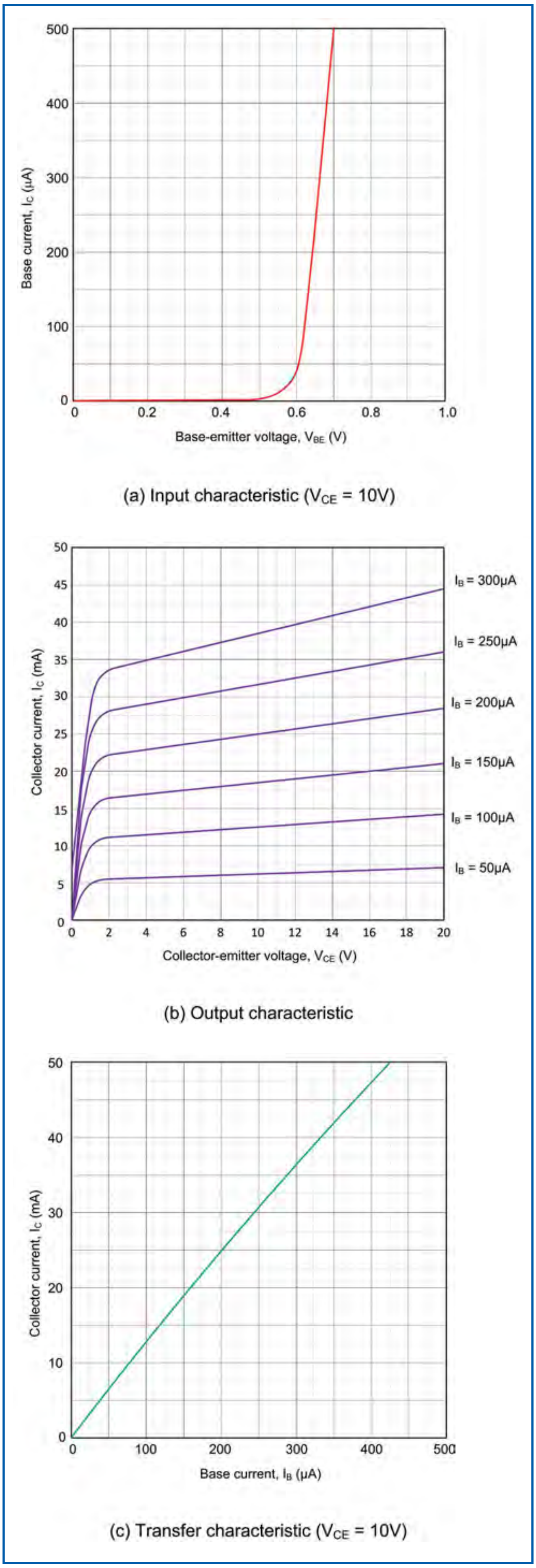


Fig.2.4 Typical characteristics for a BJT

A typical transfer characteristic (I_C plotted against I_B) for a small-signal general-purpose NPN transistor operating in common-emitter mode (see later) is

Table 2.4 General hybrid parameters for a two-port network

Parameter	Meaning	Derivation	Equation
h_i	Input resistance	Ratio of input voltage to input current with output voltage held constant	$h_i = \dfrac{\Delta V_i}{\Delta I_i}$
h_r	Reverse voltage transfer ratio	Ratio of input voltage to output voltage with input current held constant	$h_r = \dfrac{\Delta V_i}{\Delta V_o}$
h_f	Forward current transfer ratio	Ratio of output current to input current with output voltage held constant	$h_f = \dfrac{\Delta I_o}{\Delta I_i}$
h_o	Output conductance	Ratio of output current to output voltage with input current held constant	$h_o = \dfrac{\Delta I_o}{\Delta V_o}$

Table 2.5 Hybrid parameters for a common-emitter BJT amplifier

Parameter	Meaning	Derivation	Equation
h_{ie}	Input resistance	Ratio of base-emitter voltage to base current with collector-emitter voltage held constant	$h_{ie} = \dfrac{\Delta V_{be}}{\Delta I_b}$
h_{re}	Reverse voltage transfer ratio	Ratio of base-emitter voltage to collector-emitter voltage with base current held constant	$h_{re} = \dfrac{\Delta V_{be}}{\Delta V_{ce}}$
h_{fe}	Forward current transfer ratio	Ratio of collector current to base current with collector-emitter voltage held constant	$h_{fe} = \dfrac{\Delta I_c}{\Delta I_b}$
h_{oe}	Output conductance	Ratio of collector current to collector-emitter voltage with base current held constant	$h_{oe} = \dfrac{\Delta I_c}{\Delta V_{ce}}$

shown in Fig.2.4(c). This characteristic shows the almost linear relationship that exists between collector current and base current (eg, doubling the value of base current produces double the value of collector current, and so on). This characteristic is reasonably independent of the value of collector-emitter voltage (V_{CE}) and thus only a single curve is shown. Finally, it's important to note that the input and transfer characteristics are both measured with the collector-emitter voltage (V_{CE}) held constant.

The hybrid equivalent of a BJT

In last month's *Teach-In 2015* we introduced the equivalent circuit for an amplifier. This allowed us to express the overall voltage gain of the amplifier taking into account the amplifier's input and output resistance. We use a slightly more complex equivalent circuit model in the analysis of a transistor amplifier. This represents a transistor using four basic 'hybrid' parameters or *h*-parameters. The term 'hybrid' is used to describe these parameters because we use both voltage and current sources to model the transistor's behaviour.

In the general *h*-parameter equivalent circuit (see Fig.2.5) the transistor is modelled by just four components; h_i, h_r, h_f and h_o (see Table 2.4). These four components are arranged in a network which has two 'ports'; an 'input port' and an 'output port', as shown in Fig.2.5. Note that in Table 2.4 we have introduced the Greek symbol, Δ (upper-case delta), to indicate that we are dealing with an incremental change rather than steady-state DC values. This is important because we are attempting to model the amplifier's response to relatively small AC signals.

We have indicated the voltage and current present at the input and output ports of Fig.2.5. The variables at the input port are marked V_i and I_i and at the output port we have V_o and I_o. Since there are four variables, when we come to specify the *h*-parameters in terms of these variables we need to hold one of the other variables constant. For example, when we specify input resistance (V_i/I_i) we need to do this for a specified value of output voltage (V_o). We have included this information in each of the derivations shown in Table 2.4.

In order to indicate which one of the operating modes is used we add a further subscript letter to each *h*-parameter; *e* for common emitter, *b* for common base and *c* for common collector (see Table 2.5). To keep things simple, we will only consider common-emitter operation, in which case the input port is between the base and emitter and the output port is between the collector and emitter, see Fig.2.6.

Now let's look at each one of the four common-emitter hybrid parameters in turn in relation to popular BJT devices. In Fig.1.4 last month we showed a sample datasheet for the BC546, BC547, BC548 and BC549 family of NPN general-purpose transistors. The information included a set of typical hybrid parameters for devices within each gain group. It's worth taking another look at the datasheet, but for a device in Gain Group B you should note the following typical *h*-parameter values:

- h_{ie} varies between 3.3kΩ and 8.5kΩ. A value in this range is neither very high nor is it very low and we might consider it to be a medium value of resistance of, on average, about 5kΩ
- h_{re} is typically 2×10^{-4}. To put this into context, if the collector-emitter voltage increases from 4V to 5V (a change of 1V), the base-emitter voltage will experience a change of a mere 200µV. This change is very small and in most cases can be ignored
- h_{fe} is quoted as 330 typical. Hence, if the base current increases by 10µA the collector current will experience a corresponding increase of 3.3mA. This very significant increase is due to the transistor's amplifying action
- h_{oe} is stated as varying from 30µS to 60µS. Let's attempt to put this into context by taking an average value for h_{oe} of 45µS. Recalling that conductance (unit, siemen – hence 'S' or 'µS') is the inverse of resistance, so we can conclude that the device has an output resistance of around 22.22kΩ. When compared with a typical value of load resistance (see last month) this is a relatively large value (often around 10 times larger than a typical value for R_L) so we can usually ignore the effect of the output conductance unless we need to work to a very high degree of accuracy.

Now, provided that h_{re} and h_{oe} are both very small we can ignore the effects of these two parameters. This makes it possible to derive an approximate model of the BJT common-emitter

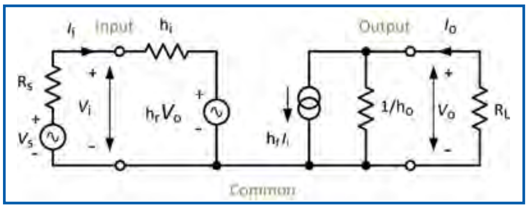


Fig.2.5 General h-parameter equivalent circuit

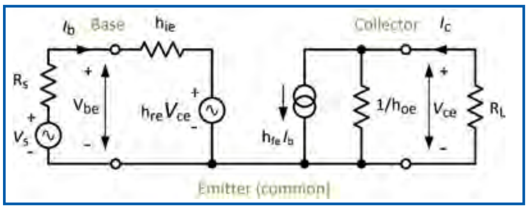


Fig.2.6 Common-emitter h-parameter equivalent circuit

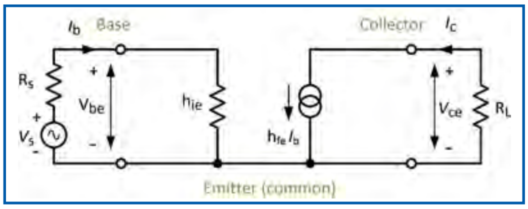


Fig.2.7 Simplified h-parameter equivalent showing the two most significant parameters

amplifier using just two of the four hybrid parameters, see Fig.2.7. In this case, the input resistance and voltage gain will be given by the two approximate relationships:

$$R_{in} \approx h_{ie} \text{ and } A_V \approx \frac{h_{fe} \times R_L}{R_s + h_{ie}}$$

where R_s is the internal resistance (impedance) of the signal voltage source. If this is very much smaller than h_{ie} we can further approximate the voltage gain relationship to:

$$A_V \approx \frac{h_{fe} \times R_L}{h_{ie}}$$

As an example, let's determine the voltage gain and input resistance of the simple common-emitter amplifier shown in Fig.2.8. In this circuit, the load resistance (R3) is 2.2kΩ and the transistor (a device from Gain Group B) has a typical h_{fe} of 330. A typical value of h_{ie} for a Gain Group B device is about 5.9kΩ. We will assume a typical value of 600Ω for R_s, in which case:

$$A_V \approx \frac{h_{fe} \times R_L}{R_s + h_{ie}}$$

$$= \frac{330 \times 2200}{600 + 5900} = 111$$

The bias potential divider chain formed by R1 and R2 will have the effect of reducing the impedance of the stage when looking into the input, but we can usually assume that, for a signal frequency in the mid-band range, the reactance of the two coupling capacitors C1 and C2, and the decoupling capacitor C3 will be very small and thus negligible. In this case, the input resistance of the amplifier stage will be the parallel combination of h_{ie} with the two bias resistors, R1 and R2. Thus:

$$\frac{1}{R_{in}} = \frac{1}{R1} + \frac{1}{R2} + \frac{1}{h_{ie}}$$

$$= \frac{1}{10{,}000} + \frac{1}{3{,}300} + \frac{1}{5{,}900}$$

$$= 0.00057$$

From which: $R_{in} \approx 1.75\text{k}\Omega$

Notice in the circuit of Fig.2.8 that the two resistors in the bias potential divider have the effect of dragging down the input resistance of the stage, which in turn, will have the effect of reducing the overall voltage gain due to the resistance of the source that drives the amplifier. Because of this we might need to ensure that the signal source has the lowest possible resistance.

Having demonstrated the use of h-parameters for predicting the performance of a simple single-stage amplifier we will henceforth leave any further analysis to the powerful SPICE packages that will perform all of the complex calculations for us. If you wish to check the pre-amplifier circuit out with TINA V10 Student Edition (see last month) we have included a sample file for downloading from the *EPE* website at: **http://www.epemag.com/projects.html**.

We need to make a very important point here; due to differences in SPICE parameters you should not expect the measured voltage gain to be the same as that obtained here. Using Tina Student Edition you should find that the voltage gain is somewhere in the range 80 to 120 which is normal for amplifiers of this type. Finally, it is important to note that, in order to cater for a wide range of transistors (recall that h_{FE} for a Gain Group B device can vary from 200 to 450), a better design would have incorporated negative feedback to define the overall voltage gain and thus compensate for real-world variations in individual transistors. We will show how this vital design stage can be achieved in a future instalment of *Teach-In 2015*.

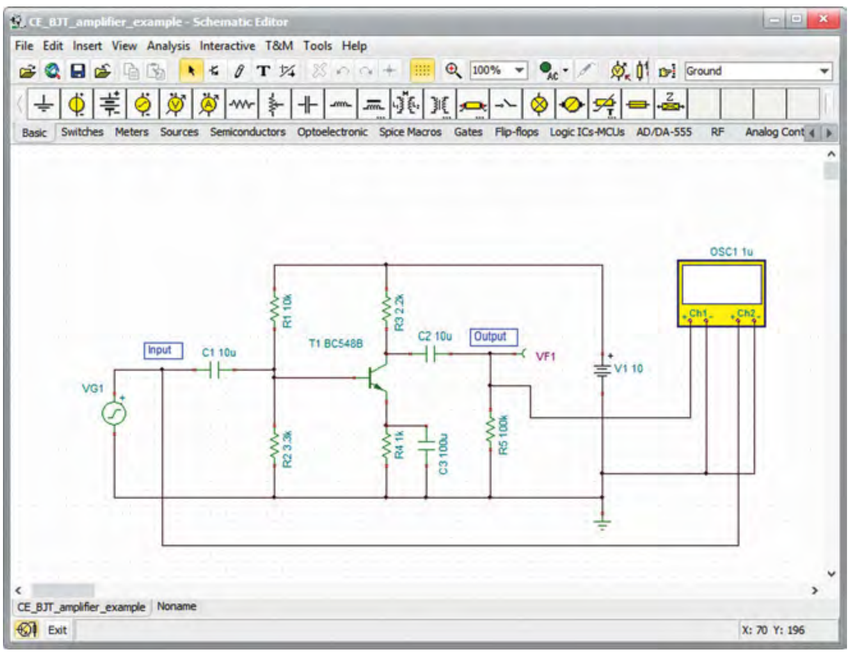


Fig.2.8 Simple common-emitter amplifier stage

Get Real: Designing and testing the pre-amplifier

Our first *Get Real* project was a simple pre-amplifier that can be used in a variety of practical applications including simple 'signal boosters', microphone pre-amplifiers, and measurement systems. The circuit was designed to produce a modest amount of voltage gain (around 25) over a wide range of frequencies extending from less than 10Hz to around 100kHz. Another important requirement was the ability to easily tailor the voltage gain and frequency response. This month we will see how close we came to achieving our design objectives and how the circuit can be easily modified for use in different applications.

Simulating the pre-amplifier

In last month's *Special Feature* we introduced TINA, a powerful design tool. It's now time to put this excellent tool to good use by showing you how we carried out an analysis of our original circuit design before we actually built it. First, you will need to enter the circuit into TINA's drawing area (or download the circuit from the *EPE* website by going to: **http://www.epemag.com/projects.html**). Check that the value of each component is set correctly and that the correct type (BC549C) has been entered for each of the two transistors. Also check that the supply voltage has been set correctly to 9V and not left at the default 5V.

Finally, click on the circuit diagram and set up signal voltage generator VG1, so that it produces a sinewave signal with an amplitude of 50mV (100mV$_{pk\text{-}pk}$). To do this click on VG1 and select 'Signal' then click on the three dots to set the amplitude to '50m' and the frequency to '1k'. Note that the default value of internal resistance will be '0' ohms. You will now be ready to carry out an analysis of the DC and AC conditions present in the circuit.

DC analysis

The first task (before we apply any signals to our circuit design) is to check that we

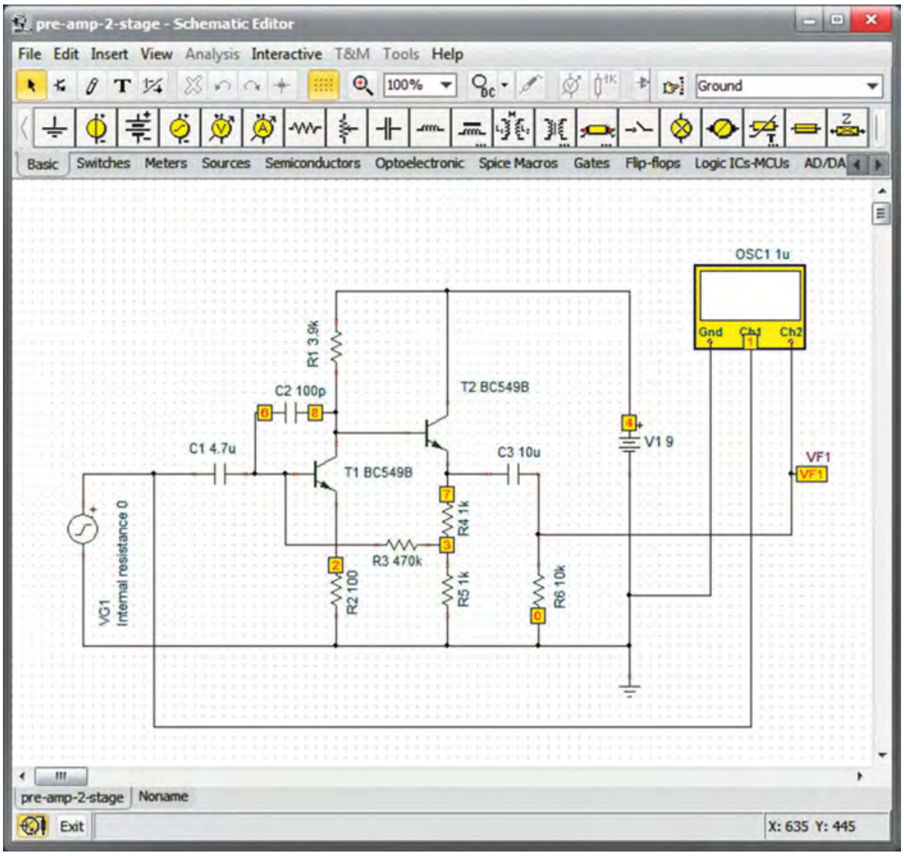


Fig.2.9 Pre-amplifier circuit (with nodes displayed) using TINA

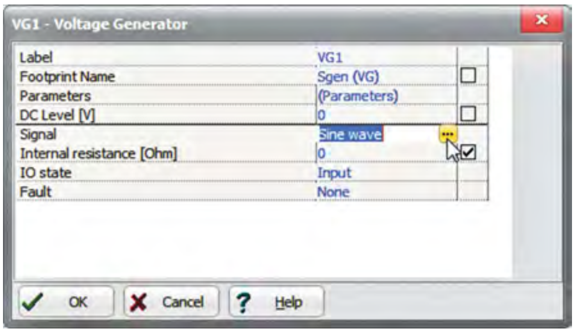


Fig.2.10 The virtual voltage generator (VG1)

have the correct DC conditions. You can do this by selecting 'Analysis' from the main menu bar at the top of the screen and then 'DC Analysis' from the drop-down list. Next, select 'Table of DC results' to produce a complete set of results for the DC currents and voltages in the circuit presented in tabular form (see Fig.2.12). Alternatively, you can select 'Calculate nodal voltages' to display the voltage at any node in the circuit, in which case the pointer will change to show a test probe and the voltage at the node will appear in a window (see Fig.2.13).

AC analysis

Next, you should carry out an AC analysis of the pre-amplifier. This will let you know how the pre-amplifier will perform when a signal is applied. You can do this by selecting 'Analysis' from the main menu bar at the top of the screen and then 'AC Analysis' from the drop-down list. Now select 'Table of AC results' to produce a complete set of results for the AC currents and voltages in the circuit presented in tabular form. Alternatively, you can select 'Calculate nodal voltages' to display the voltage at any node in the circuit, in which case the pointer will change to show a test probe and the voltage at the node will appear in a window, as before.

Signal analysis with TINA's virtual oscilloscope

Next select 'T&M' (Test and Measurement) from the menu bar at the top of the main screen and then 'Oscilloscope' from the drop-down menu. When the virtual oscilloscope opens set OSC1_Ch1 (Channel 1) to 100mV per division and OSC1_Ch2 (Channel 2) to 500mV per division, as shown in Fig.2.14.

Frequency response

To investigate frequency response, once again select T&M from the menu bar and then choose 'Signal Analyzer' from the drop-down list of virtual test instruments. Set 'Display' to Lin; Magnitude with a High of '30' and a low of '0' (this selects a linear vertical amplitude scale over the range 0 to 30. Next, choose 'Set Mode' and select Swept-sine with Start and Stop frequencies of '1' (1Hz) and '1m' (1MHz) respectively. This will set the horizontal scale to occupy a frequency range extending from 1Hz to 1MHz (six logarithmic cycles). Finally, click on Start to display the pre-amplifier's frequency response. You should find that the pre-amplifier has a frequency response that is substantially flat from about to 10Hz to 100kHz.

To produce a more detailed frequency-response graph press the right Data button and a new window will appear with the response curve (see Fig.2.16). When the detailed response is displayed it is possible to move the cross hair cursor across the frequency response and the display in the status bar (at the bottom of the main TINA window) will show corresponding values of frequency and gain.

Effect of signal source impedance

As mentioned earlier in *Knowledge Base*, the internal resistance (impedance) of the signal source can have a significant effect on the overall voltage gain of the amplifier. It can also have a considerable impact on the frequency response of the amplifier. During the previous measurements we had assumed that the signal source was perfect and had used the default voltage source resistance

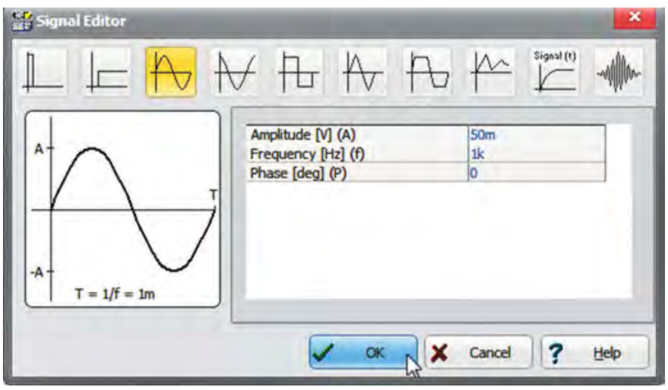
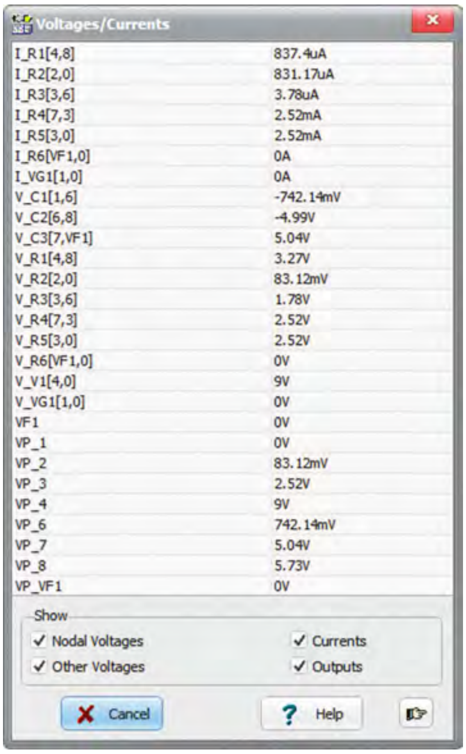


Fig.2.11 Setting waveform, amplitude, frequency of VG1


Voltages/Currents

I_R1[4,8]	837.4uA
I_R2[2,0]	831.17uA
I_R3[3,6]	3.78uA
I_R4[7,3]	2.52mA
I_R5[3,0]	2.52mA
I_R6[VF1,0]	0A
I_VG1[1,0]	0A
V_C1[1,6]	-742.14mV
V_C2[6,8]	-4.99V
V_C3[7,VF1]	5.04V
V_R1[4,8]	3.27V
V_R2[2,0]	83.12mV
V_R3[3,6]	1.78V
V_R4[7,3]	2.52V
V_R5[3,0]	2.52V
V_R6[VF1,0]	0V
V_V1[4,0]	9V
V_VG1[1,0]	0V
VF1	0V
VP_1	0V
VP_2	83.12mV
VP_3	2.52V
VP_4	9V
VP_6	742.14mV
VP_7	5.04V
VP_8	5.73V
VP_VF1	0V

Show
✓ Nodal Voltages ✓ Currents
✓ Other Voltages ✓ Outputs
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Fig.2.12 Table of DC analysis results

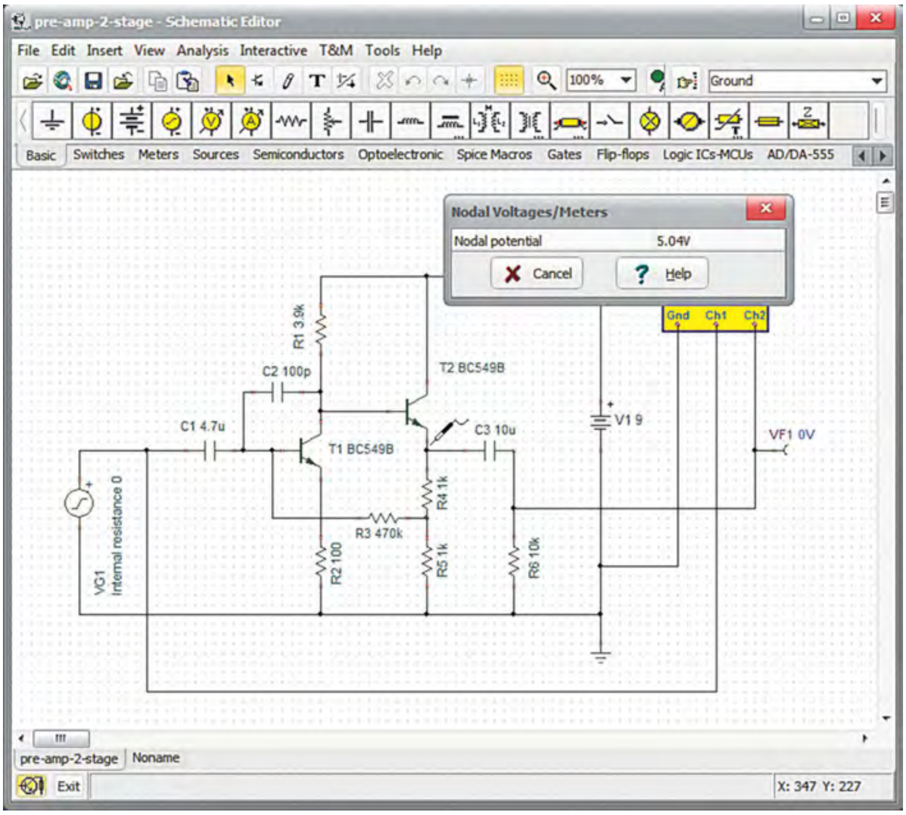


Fig.2.13 Using a virtual test probe (displaying the voltage at the emitter of T2)

of zero ohms. Increasing the source resistance to a more realistic value of 600Ω has the effect of reducing the mid-band voltage gain to 27 (not much change) but reducing the upper cut-off frequency to about 80kHz (from around 400kHz). A further increase to 5kΩ produces a mid-band voltage gain of 21 (below the target value of 25) and an upper cut-off frequency of a mere 14kHz. However, all is not lost because we can reduce the value of shunt voltage feedback capacitor C2, to compensate. With C2 at 22pF (instead of the original 100pF) the upper cut-off frequency is increased to around 60kHz. With C2 at 33pF the upper cut-off is 40kHz, making this value acceptable for good quality audio applications (there's no point in an amplifier having a frequency response that extends way beyond the highest frequency signal component). Note from all of this that the value of C2 and the signal source resistance both have a major impact on the high frequency performance of the pre-amplifier.

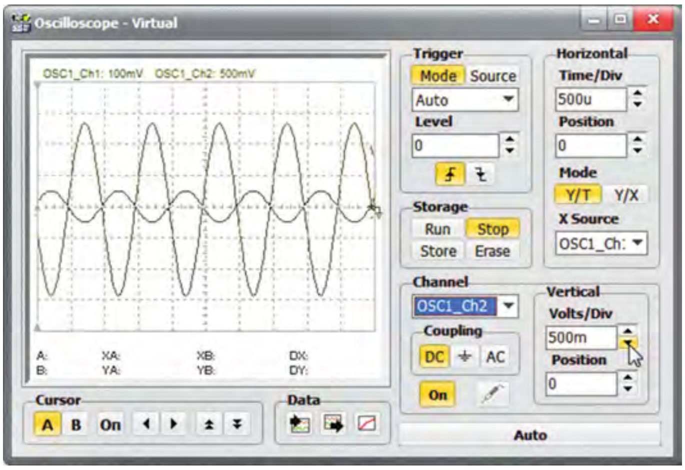


Fig.2.14 TINA's virtual oscilloscope displaying input and output signal voltages for the pre-amplifier

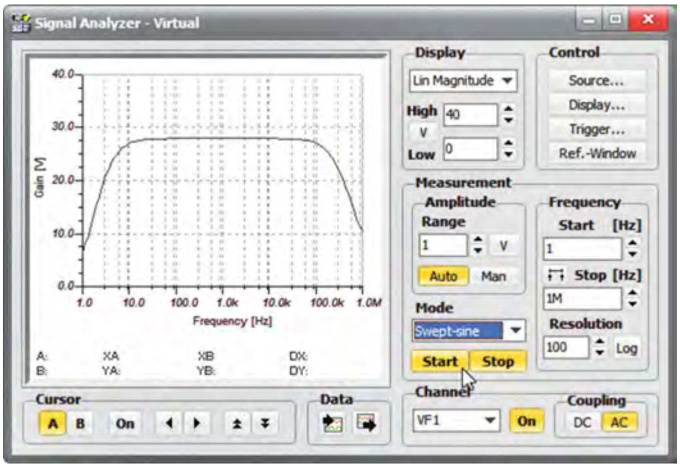


Fig.2.15 The frequency response of the pre-amplifier extends from about 10Hz to 100kHz

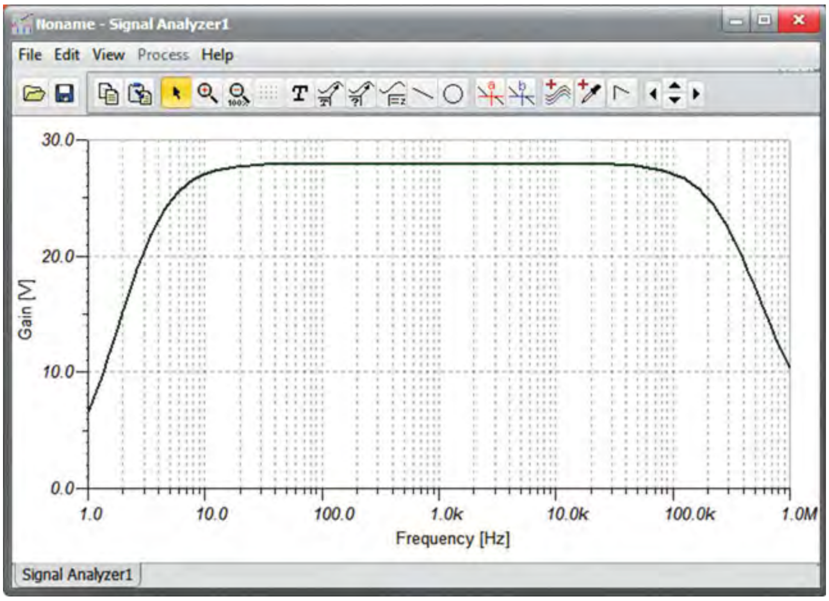


Fig.2.16 A more detailed frequency response via the Data button

Effect of series current negative feedback

By introducing a resistor (R2) in the emitter circuit of TR1 we have introduced series current negative feedback. The larger the value of this resistance the greater the amount of feedback and, as a consequence, the less the overall voltage gain will be. The value of R2 cannot be increased to more than about 400Ω without having an effect on the DC bias conditions and hence we have suggested an acceptable range of values for R2 from zero (short-circuit) to 330Ω. Note that when the value of R2 is increased the input resistance of the amplifier will also be increased by virtue of the series current negative feedback. Table 2.6 shows how the pre-amplifier can be configured for various applications by means of changes to R2 and/or C2.

Testing the pre-amplifier

Having completed our TINA simulation of the pre-amplifier we were reasonably satisfied that the circuit would operate according to our original design specification (see last month's *Teach-In 2015*). We then went on to develop

Table 2.6 Values of mid-band voltage gain and upper cut-off frequency for different combinations of R2 and C1 (signal generator output impedance = 600Ω)

R2	C1	Mid-band voltage gain (1kHz)	High-freq cut-off	Application
zero	47p	94	54kHz	High-gain general purpose audio
zero	100p	94	25kHz	High-gain microphone pre-amplifier
47Ω	100p	43	50kHz	Reduced gain general purpose audio
100Ω	47p	27	150kHz	High-quality audio
220Ω	100p	15	120kHz	Wide-band 'signal-booster'
330Ω	100p	10	117kHz	×10 amplifier with approx. 50kΩ input

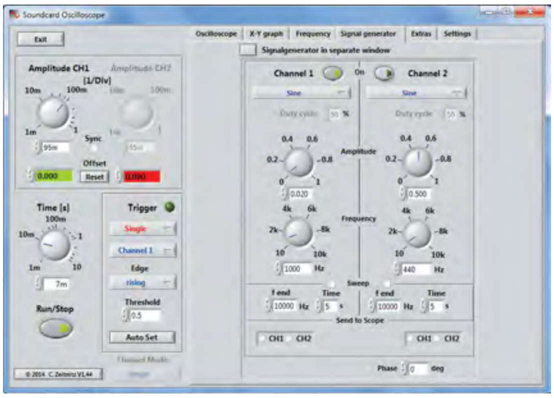


Fig.2.17 The Soundcard Scope signal generator window with Channel 1 output selected for use as a signal source for testing the pre-amplifier. Note that 'sine' has been selected, the output voltage has been set to '0.02' (20mV) and the output frequency to '1000' (1kHz)

a printed circuit board layout (see our *Special Feature* next month) before building the circuit and testing it using a variety of 'real' and virtual test instruments (we used the 'real' test instruments in our own test lab to check the accuracy of the measurements that we obtained from the virtual instruments).

We used the two virtual instruments mentioned earlier (Soundcard Scope and Multi-instrument from Virtins) to carry out an extensive range of measurements on the prototype pre-amplifier and to check that the measured performance was consistent with the results that we obtained from the TINA simulation. If you have built your own version of the pre-amplifier we suggest that you carry out your own similar measurements, starting with a straightforward DC voltage check using a standard bench multi-meter in order to verify that you have the correct bias potentials for the two transistors. We have shown our results in Table 2.7 (yours should be within ±10% of those shown here).

Having completed your DC check, the next stage will be that of connecting a test signal to the input of the pre-amplifier. The procedure is different for the two virtual instruments but both of them incorporate a signal generator and you will need to set this so that it produces a sinewave signal at 1kHz with an amplitude of 20mV (see Fig.2.17). The corresponding input and output waveforms are shown in Fig.2.18. Finally, Fig.2.19 shows the effect of over-driving the input of the pre-amplifier. Notice how the waveform has become severely clipped on both positive and negative peaks.

We have summarised all of our simulated and measured results in Table 2.8. These show that there is fairly close agreement between the test lab and simulated measurements and that, in most cases, we managed to achieve results that were in line with the original design objective. In a future instalment of *Teach-In 2015* we will show you how to carry out a variety of more sophisticated measurements (including bandwidth, distortion, input and output impedance) using virtual instruments.

Next month

In next month's *Teach-In 2015*, *Get Real* will feature the design and construction of a simple headphone amplifier, *Discover* will introduce you to power amplifiers and *Knowledge Base* will explain classes of operation. For good measure, our *Special Feature* will show you how you can use Circuit Wizard to design your own printed circuit boards.

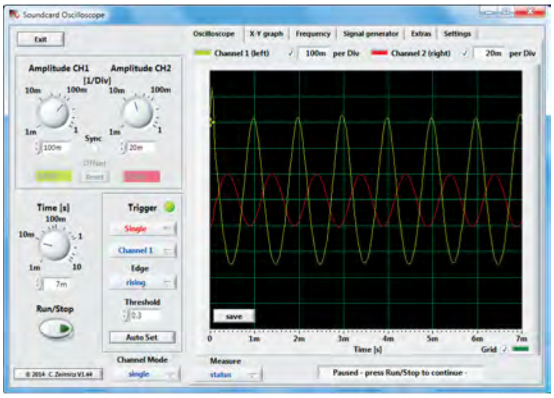


Fig.2.18 The Soundcard Scope oscilloscope window showing the input (red) and output (green) waveforms. Note that the vertical scale has been set to 100mV per division for the output waveform and 20mV for the input waveform. You might like to compare this display with that obtained during simulation (Fig.2.14)

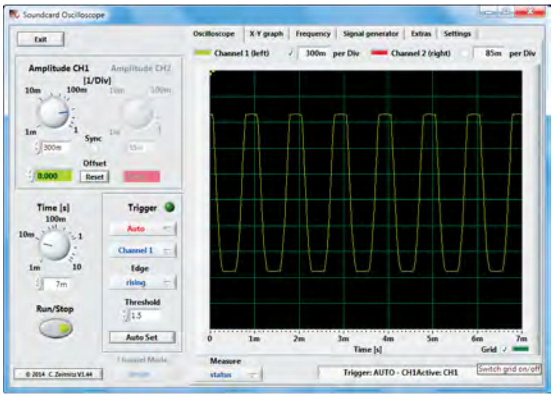


Fig.2.19 Effect of over-driving the input of the pre-amplifier on the output waveform

Table 2.7 Measured (M) and simulated (S) bias conditions for TR1, TR2 (differences due to a slightly higher supply voltage from a 'fresh' PP3 battery and some variation in transistor parameters)

Simulated	Base		Emitter		Collector	
	S	M	S	M	S	M
TR1 (T1)	0.74V	0.75V	0.08V	0.1V	5.73V	5.41V
TR2 (T2)	5.73V	5.41V	5.04V	4.74V	9V	9.3V

Specification	Design objective	Simulated (TINA)	Measured (virtual instrument)	Measured (test lab)
Mid-band voltage gain (1kHz)	>25	30	30	30
Input impedance (1kHz)	20kΩ	20kΩ	18kΩ	18kΩ
Output impedance (1kHz)	30Ω	25Ω	25Ω	27Ω
Lower cut-off frequency	10Hz	2.5Hz	3Hz	2.8Hz
Upper cut-off frequency	100kHz	70kHz	65kHz	68kHz
Noise (ref. 1V in 100kHz bandwidth)	Better than –65dB	Not measured	Not measured	–72dB

Table 2.8 Comparison of measured and simulated circuit performance

NET WORK

by Alan Winstanley

Lifting the lid of your network

LAST MONTH, I suggested a way of restoring the Network Activity Icon to Windows, a visual indicator last seen in Windows XP. The flashing monitor icon resides in the system tray and shows the PC is actively communicating on the network. A modern replacement can be downloaded free from: **www.itsamples.com**.

Monitoring behind the scenes activity

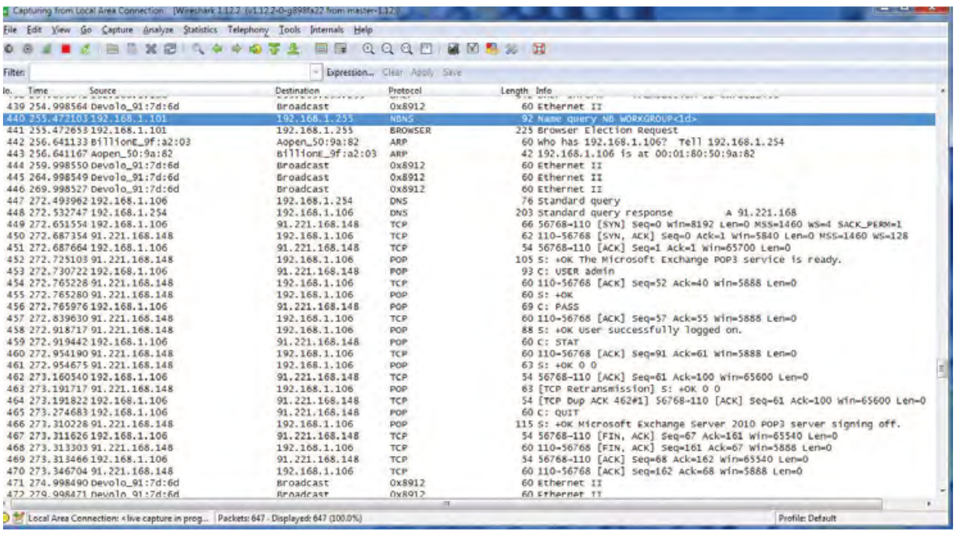
Obviously, this only provides a simple on-off indication of network activity. However, an awful lot can be going on in the background and readers can be forgiven for wondering what their PC is sometimes doing, especially when a lot of disk activity is evident after booting up. Software such as Adobe Reader, Flash Player or AIR may have new updates queuing up ready to install, or an anti-virus program or Windows itself may be limbering up ready to hit your system with an avalanche of updates. They may also be stacked up from previous sessions waiting to be installed. (Some users complain that seldom-used laptops are so slow to start that they never bother with them, but often if left alone for 10-20 minutes to catch up with updates, boot-up times should improve.) Worst case, a system may be hosting unwanted Trojan or malware programs that generates suspicious Internet traffic without your knowledge.

Internet traffic consists of packets of data transported between IP addresses. In essence, multiple layers of protocols are contained within TCP/IP (Transmission Control Protocol/ Internet Protocol), the common language spoken by Internet-connected devices and other networks. The traffic itself consists of data packets handled by the Internet and 'network access layers', with TCP providing the 'transport layer' – the rails that traffic runs on. Various error-checking routines ensure that data is sent and received reliably between IP addresses.

If users want to see what their computer is actually 'doing', then some readily available software will provide an insight into their network traffic. The program **Wireshark** is a free and sophisticated analyser that logs traffic on your Windows PC, Linux or Mac OS X system (10.6+). It can be downloaded free from: **https://www.wireshark.org**. Current versions support Windows Vista onwards; legacy XP users should use V1.10 only. After installing Wireshark, point it to your computer's network interface card (NIC) that carries your IP traffic, and in the Capture pane menu (press CTRL + I) select your machine's network interface (eg, 'Local Area Connection'). In my case, a Bluetooth network was also shown, so ensure that the Internet network connection is selected, then click the green Start icon. (In View, just tick 'Packet List' for a basic view.) Also enable 'Auto Scroll in Live Capture' in the View menu and you can colour-code different types of event (View/Colouring Rules).

What does it all mean?

Various columns show source and destinations of packet traffic, where an IP address or ID will also be seen. Your screen will reveal the real extent of your computer's network activity, with all manner of traffic passing through your network interface, including Homeplug, NAS and traffic to unrecognisable third parties. The next question is, what do all those IP addresses mean? Some of them are local devices residing on your network, but others are outside on the WAN (wide-area network) instead. Using online tools such as **dnsstuff.com**, an unknown IP address can be looked up and translated into a more familiar form, perhaps a vendor's name and address offering a clue. One online service – Dropbox – was seen as causing much network activity during booting up when it syncs my Dropbox folders. Links to Avast anti virus server IPs were also spotted as my PC reached out onto the web, and simply by doing a WHOIS lookup on an IP address the vendor's name was established.



Wireshark offers users a glimpse of traffic passing over a system's network

Wireshark sessions can be saved to a file for future reference. The software is also handy for helping diagnose strange problems, such as email login failures, by clearly showing any error messages that might arise during handshaking; for example, it clearly highlighted how my email was failing to log in properly using authenticated SMTP at one stage. In fact, this was a self-inflicted problem caused by changing a POP3 (incoming) mailbox login, which broke the SMTP (outgoing) authentication rules and caused a corresponding failure when *sending* mail.

Command and control

Windows isn't always the answer to every network problem and there

are times when a command-line approach is faster and more powerful. Readers can try some DOS-style commands by typing **CMD** in the Windows Search box, then type `ipconfig /all` for a summary of your network setup. You can also use Ping to check that your broadband connection is up and running at grass roots level. For example, type: `ping www.ebay.com`, and eBay's IP address will be looked up, which at least proves you have an Internet connection because your ISP's DNS servers (which translate URLs into IP addresses) are working. Last, at the DOS prompt, type: `netstat /?`, which gives you a summary of the options available for the netstat command (note that `/?` works on any command). If you want to experiment or gain an insight into your network then Wireshark is a useful tool to keep in a surfer's armoury, so try checking your email or surfing the web when Wireshark is running, and see the traffic flowing over your network in real time.

A book in all of us

For decades, many have predicted how technology would revolutionise our lives in the future. The fascinating 1967 movie, called *1999 AD* and made by America's Philco-Ford Corp, speculated about the impact that futuristic 'home computers' might have in the home – it can be viewed at: **https://www.youtube.com/watch?v=TAELQX7EvPo**. Although some ideas were far-fetched, flat-screen computers, large video displays, home information servers, music synthesisers, online shopping and e-commerce, networked security cameras and the equivalents of YouTube and Skype were all foreseen by Philco (now part of Philips) as the home computer promised to make our 'space-age dreams come true'.

Fast forward to the 1980s, and it was reckoned that home computers could control our homes and even dispense with the need for text books: cooks toiling in the kitchen were pictured keying into a Sinclair Spectrum computer instead of thumbing through a recipe book. Thirty years later, and some of those wild speculations are with us as recipes can be downloaded on to tablets, smartphones and even via some Internet-aware kitchen appliances. Apps help us to control our homes and entertain us on demand. Interestingly, tablet sales almost plateaued last Christmas compared to 2013 because it seems virtually everyone now owns one.

In the 21st Century, reading a newspaper or PDF with a tablet while on the train or beach is now an everyday occurrence, but books and periodicals printed on a traditional material called 'paper' are still in great demand: after all, there is still no substitute for the tactile feeling of thumbing through a favourite magazine or immersing oneself in a novel over coffee.

Some online publishing services also provide budding authors and publishers with practical ways of turning their ideas into reality. Far from causing the demise of printed media, the Internet is proving to be a blessing in disguise for emerging publishers or writers, as it has opened up the prospect of self-publishing one's work at minimal cost and reaching a worldwide audience at the same time.

Websites such as Blurb and Lulu (**lulu.com** and **blurb.co.uk**) allow individuals and small publishers to upload their work as a PDF, and books can then be printed and bound to order for shipping worldwide; PDF versions can also be sold. It may sound hard to believe, but 'print on demand' services like these offer high quality reproduction with full colour digital printing and binding, producing books and magazines to order – whether a single copy or a thousand – that are almost indistinguishable from those printed using traditional offset-printing methods.

Up the Amazon

Think of online book sales and one immediately thinks of Amazon. There is no denying the formidable, world-class marketing machine that Amazon has become over the past 20 years, but Amazon also offers potential writers ways of entering both Kindle ebook and traditional book markets for themselves with no startup costs. Amazon Kindle ebook reader software is freely available for Windows, Mac and smartphones, so ebook sales are not restricted to Kindle tablet owners; virtually anyone with an Internet connection can enjoy reading Kindle books and Amazon Prime users can borrow them free via the Kindle Owners' Lending Library (KOLL).

Whether the topic is non-fiction, sharing some resources for fellow hobbyists or writing a life history or children's book, Amazon Kindle offers a very good foundation for emerging writers and there is plenty of support available to help them get going. The place to start is Kindle Direct Publishing (KDP) at: **https://kdp.amazon.com**. Note this is only available on the US website, from where Kindle ebooks originate. This is not a problem, but do remember that UK readers, for example, can only download a Kindle ebook from the Amazon UK website. The same is true for corresponding overseas Amazon sites.

Publishing

After creating an account, a 'New Title' can be added, which is the start of the Kindle publishing process. A Kindle ebook's sales price is set by its author, and authors receive royalties on sales depending on the selected market and publishing model, and are further subject to US Withholding Tax of 30%. It is enormously complicated for foreign writers to claw back this tax and it may be more cost-effective just to swallow the deduction rather than spend an awful lot of time applying for an IRS ITIN number. Authors should study Amazon's KDP pricing page and Terms and Conditions closely.

Writing

The main focus of producing a Kindle ebook is of course the actual writing and composition. For many writers,

Online shopping predicted in Philco-Ford's 1967 educational film 1999AD

1999AD: Flat screen monitors and video communications provided by a home computer. A user chats and checks the weather forecast 'online'. (Philco-Ford, 1967)

Microsoft Word offers a satisfactory solution for text-only work and an entire Kindle book can be drafted using this program: Kindle's system understands Word very well. Note that different Kindles have different screen sizes, giving users the ability to zoom in or out; this makes the book's flow of text and page breaks important if the Kindle user is to enjoy a seamless 'read' without any jarring on-screen interruptions to the pages. An initial ToC (Table of Contents) with hyperlinks to each chapter is also needed, something that Word will readily create. A cover graphic can also be produced to accompany your Kindle book, and this will appear in Amazon's websites worldwide.

Images

If your work needs illustrating then embedding images and captions into text can be more of a problem, as Microsoft Word makes a hash of things. In the past, I have found that photos were churned out as GIFs, while logos and line art (ideal for GIFs) were converted into JPEGs. Word has no workaround for this, and more advanced software users would be better creating the final file in an HTML program like Dreamweaver. This will ensure JPEGs are embedded properly to produce an all-important HTML file with a separate folder of hyperlinked images, or, more advanced still, use Adobe InDesign, which is really aimed at serious designers.

Amazon tools

Amazon offers free previewing and conversion tools, which should be used for testing prior to uploading a completed file into your KDP account. Kindle will then process the entire uploaded project and images into a suitable format and add it to your account. Other formalities, such as a simple online tax interview should also be completed.

A lot of online help is available in the KDP website, which will answer all of the above aspects. My own advice is to research the technical and commercial aspects thoroughly to get a good all-round feel for what is required, which includes having a realistic idea of prices, markets and royalties, but don't over-analyse things. Then spend time working on the manuscript, experimenting and previewing for as long as necessary prior to uploading it to your account. After review, your Kindle ebook will go live worldwide within 48 hours and then you can start to watch royalties trickle in. Payments are made some 60 days in arrears direct to a bank account, although Amazon's online accounting system is tortuous to work with. Why not write your own ebook and give Amazon Kindle a go?

Next month, I offer some pointers towards online print-on-demand services, highlighting some sources of high-quality printed books and journals that will print and ship copies of your publication almost anywhere in the world. You can email the author at: **alan@epemag.demon.co.uk**

Practically Speaking

Robert Penfold looks at the Techniques of Actually Doing it!

Colour codes

IT is a reasonable assumption that the readers of *EPE* are familiar with units such metres, grams, and degrees Celsius, since they are the type of thing that are used regularly in everyday life. However, prior to an interest in electronics most people probably will not have encountered units such as farads, henries or ohms. Matters are complicated by the fact that a very broad range of values are often used. A further difficulty is that some components are not labelled with a value, but instead have it marked using a system of colour coding.

Little and large

The highest value capacitors in common use have values that are more than a billion times higher than those of the lowest value types. Any electronics component catalogue should list resistors in a range of values from less than one ohm to ten million ohms or more. One slight complication is that, by normal electronic standards anyway, the basic units are often very large or extremely small. In the case of resistors you are more likely to use one having a resistance of thousands or millions of ohms than one having a resistance of just a few ohms. Things are the other way around with capacitors, where you are unlikely to encounter a one-farad component, but might often use components having values of a few billionths of a farad.

Rather than using very large or small figures for these awkward values, prefixes are used to indicate that the values are in millions of ohms, millionths of a farad, or whatever. These prefixes follow the standard metric system and are not specific to the world of electronics. For example, a kilometre is a thousand metres, and a kilohm is a thousand ohms. Table 1 shows the prefixes that are likely to be encountered in electronics, together with the single letter abbreviation used for each one. It is important to realise that case is important with these abbreviations. In particular, an upper case 'M' means a million, whereas a lower case 'm' means a thousandth. The single letter abbreviation for micro is the Greek letter mu (µ), but 'u' is often used instead.

Table 1

Multiply/divide	Prefix	Letter
x1000000000	giga	G
x1000000	meg or mega	M
x1000	kilo	k
/1000	milli	m
/1000000	micro	µ (or u)
/1000000000	nano	n
/1000000000000	pico	p

Resistance

The 'ohm' – symbol, upper case Greek letter omega 'Ω' – is the basic unit of electrical resistance. An ohm is a very small basic unit, so resistors are generally available with values from about 1Ω to 10,000,000Ω (10 million ohms), but the higher *power* types typically start at 0.1Ω and go up to a maximum of a few hundred ohms. Components having resistances of 100 million ohms or more are produced, but are difficult to use and do not feature significantly in real-world electronic circuits.

While 'Ω' is used to indicate that a value is in ohms – a 680Ω resistor has a value of 680 ohms – the letter 'R' is often used in place of omega, and a value of 680Ω might therefore appear on a circuit as either '680Ω' or '680R', or possibly even as just '680'. On circuit diagrams the character denoting the unit of measurement is often used to indicate the position of the decimal point as well. This method now seems to be in general use, and can be used in components lists and elsewhere. A 4.7-ohm resistor for example, would normally have its value written in the form '4Ω7' or '4R7'.

Because the ohm is quite a small basic unit, most real-world resistors have their values expressed in thousands or millions of ohms – kilohms or megohms respectively. The abbreviation for kilohm is 'kΩ' or just 'k', and the abbreviation for megohm is 'MΩ' or just 'M'. As with low values, the letter indicating the unit of measurement is often used to indicate the position of the decimal point as well. A value of 3.9 kilohms would normally be written in the form '3k9', and a value of 8.2 megohms would be given as '8M2'.

Colour coding

Higher power resistors mostly have the value written on the body, probably together with a tolerance rating or code letter, and the power rating. This method is sometimes used for low-power resistors, but a system of colour coding is the norm for resistors having a power rating of less than about one watt. Since a number of small resistors feature in practically every electronic circuit, it is important for beginners to become familiar with this coding as soon as possible.

There is more than one version of resistor colour coding, but the four-band version

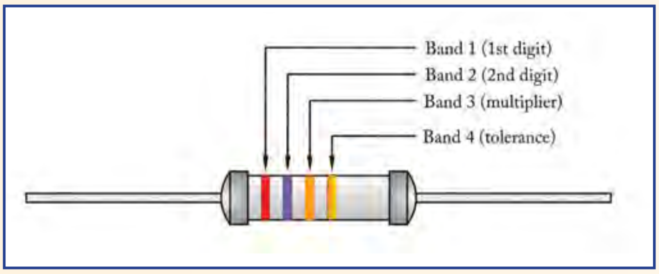


Fig.1. The standard four-band method of resistor colour coding – in this example the component has a value of 27 kilohms (27kΩ) and a tolerance rating of five per cent (ie, it's value is ±5%)

is the most common type, and the others are really just variations on the four-band system. Fig.1 shows the way in which this method of coding works, and Table 2 shows the meaning of each colour. Exactly what a colour represents, if anything, depends on the band in which it appears.

Table 2

Colour	Band1/2	Band 3	Band 4
Black	0	x1	-
Brown	1	x10	1%
Red	2	x100	2%
Orange	3	x1000	-
Yellow	4	x10000	-
Green	5	x100000	0.5%
Blue	6	x1000000	0.25%
Violet	7	-	0.1%
Grey	8	-	-
White	9	-	-
Gold	-	0.1	5%
Silver	-	0.01	10%
None	-	-	20%

In order to decipher a resistor code you first have to determine which of the bands is band one. Band 4 always used to be well separated from the other three bands, making it unlikely that bands one and four would be confused, and the code would be read in reverse. These days there is not always any obvious difference in the spacing of the four bands, but it still unlikely that a code would be read backwards. Band one is positioned close to its end of the body and in some cases it is positioned right at one end of the body. There should be no problem with the small resistors used in most projects, which have a tolerance rating of 5 per cent – this is represented by a gold fourth band, and gold is never used in bands one or two. The same is not true of close tolerance resistors having ratings of one per cent (brown) and two per cent (red), and extra care should be taken when dealing with these.

Fig.1 shows an example of a four-band colour code. Here the first two bands are red and violet, and these give the first two digits of the value. These are '2' and '7' respectively, giving a value that starts '27'. The third band is the multiplier, and the colour here is orange. This means that the first two digits must be multiplied by one thousand, or three zeros must be added to them. This gives a value of 27000 ohms, or in other words 27kΩ. The gold fourth band shows that the component has a tolerance rating of plus or minus five per cent (±5%), and that its actual value could therefore be anything from 25650 ohms (95% of 27k) to 28350 ohms (105% of 27k).

Over-specifying a component

It is perhaps worth mentioning that it is perfectly acceptable to use a component that has a tighter tolerance rating than the one called for in a components list. For instance, a component having a one or two per cent tolerance rating could be used where an ordinary five per cent type is specified. It could be an expensive way of doing things, but it will work fine. On the other hand, using a 'bog standard' component where a close tolerance type is called for is unlikely to give good results. The performance of the circuit is likely to be impaired, and it might even fail to work.

From the electrical point of view, it is also perfectly acceptable to use a resistor having a higher power rating than the one specified in the components list. However, a higher power rating is usually accompanied by a corresponding increase in physical size. Since modern circuit boards usually have very little unoccupied space, resistors of a higher power rating might not fit into the available space on the board.

Alternative bands

Rather unhelpfully perhaps, the four-band method of colour coding is not the only resistor colour code system currently in use. There are *two* five-band methods, and these operate in the manner shown in Fig.2 and Fig.3. The type shown in Fig.2 is easy to deal with, because it is essentially the same as the four-band type. In fact, the first four bands provide the value and tolerance ratings in the standard four-band fashion. The only difference is that there is a fifth band that

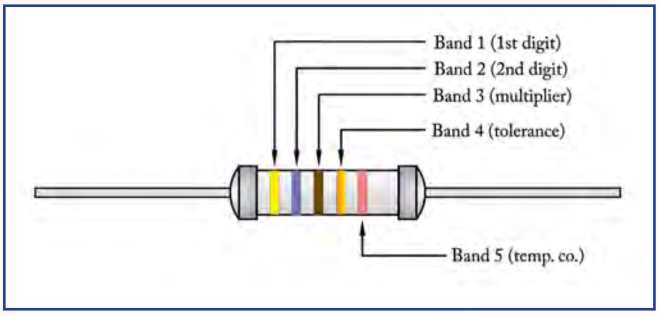
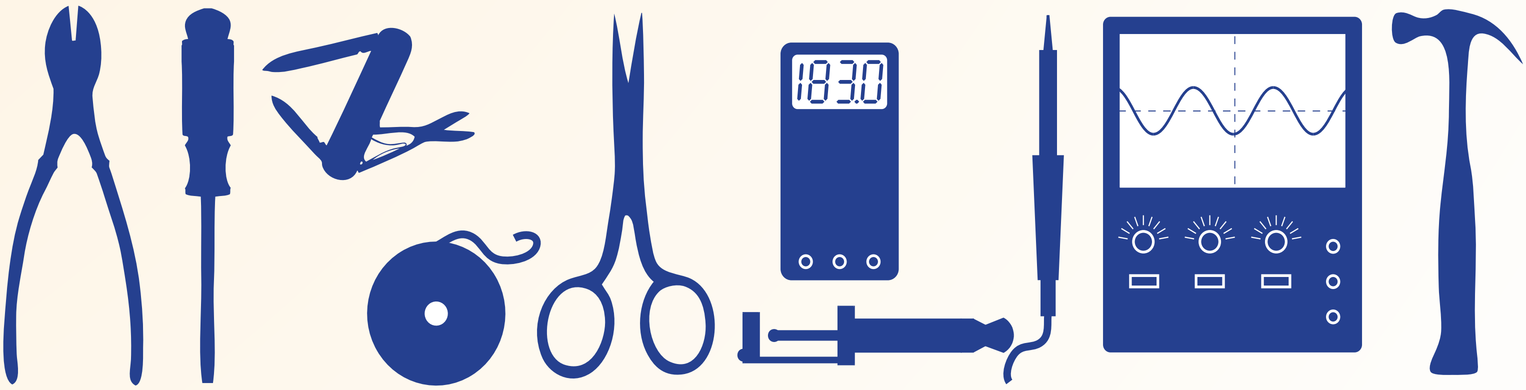


Fig.2. This five-band method of colour coding is basically the same as the four-band variety. Just ignore the fifth band and read the value in the normal way (470Ω ±5% in this example)



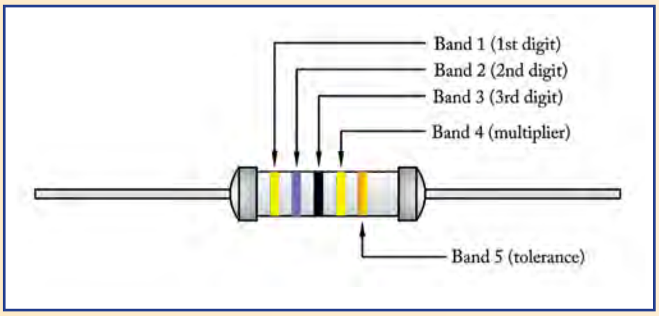


Fig.3. An alternative form of five-band colour coding that can handle an extra digit in the value. However, with 'preferred values' the third band is always black (0), and in this example the value is 4.7 megohms with a tolerance rating of five per cent (4.7MΩ or 4M7 ±5%)

indicates the temperature coefficient of the component. In other words, it indicates how much the value changes with variations in temperature. This is not something that is normally of any practical significance, and it is therefore just a matter of reading the value in the normal way and ignoring the fifth band. In this example the first two digits are '47' (yellow and violet), and the multiplier is ten (brown), giving a value of 470Ω.

The other five-band system is a bit trickier, since it does involve a slight change to the way in which the value is decoded. The four-band method of coding is adequate for handling normal resistor values, but the five-band method of Fig.3 is more versatile and can accommodate non-standard values that are beyond the normal method of coding. It achieves this by having the first three bands provide the first three digits of the value. The fourth and fifth bands then provide the multiplier and tolerance values in the normal way.

In the example of Fig.3 the first three bands are yellow, violet and black, which correspond to the digits '4', '7', and '0' respectively. The first three digits of the value are therefore '470'. The yellow fourth band indicates a multiplier value of ten thousand, or that four zeros should be appended to the first three digits. This gives a total value of 4700000 ohms, or 4.7MΩ. The gold fifth band indicates that the component has a tolerance rating of plus or minus five per cent.

This method of coding is a bit 'over the top' for the standard range of values, or 'preferred values' as they are normally termed. With preferred values the third band is not really needed and will always be black (0). The value can therefore be calculated in the normal way, but with the third band being ignored, and the decoded value being multiplied by ten in order to compensate for the ignored zero. If decoding three bands produced a figure of (say) 15kΩ, multiplying this ten would give the true value of 150kΩ.

Life will certainly be more difficult should you end up with a stock of resistors having a mixture of three types of colour codes. Finding a component of the required value from the spares box becomes more difficult, and mistakes are more likely. In particular, the type that uses four bands to carry the value does not coalesce well with the two types that use just three bands, and if possible, the five-band type of Fig.3 should be avoided. It is perhaps worth mentioning that there is even a six-band code that provides both the extra digit in the value and the temperature coefficient. Fortunately, this method of coding seems to be a rarity!

Choked up

Although colour coding is only used to a large extent with resistors, it is also used for some other types of components. Inductors, which are also known as chokes, mostly have the value, tolerance and possibly other parameters simply written on the components, but colour coding is sometimes used with the smaller types. Inductors are not used a great deal in electronic projects and are something of a rarity in comparison to resistors. The basic unit of inductance is the 'henry' – symbol 'H' – which is a huge amount of inductance, and most real-world inductors therefore have their values expressed in millihenries (mH), microhenries (µH) or nanohenries (nH). These are respectively millionths, thousandths, and billionths of a henry.

Colour coding for inductors operates in essentially the same manner as the standard four-band resistor coding, but the value is in nanohenries rather than ohms. Decoded values are divided by one thousand or one million respectively in order to give an answer in microhenries or millihenries. If the component in Fig.1 was an inductor rather than a resistor, its value would be 27000nH, or 27µH.

Fig.4. Colour codes are not used a great deal with capacitors these days. The capacitor shown here is an old 2.2 microfarad component with a tolerance of plus or minus ten per cent (2µ2 ±10%), and a maximum operating voltage of 250V

Capacitors

Although colour coding was once common for various types of capacitor, and it does have some advantages, it no longer seems to be used to a significant extent with capacitors. If you come across a colour-coded capacitor it should be easy to decipher its value. The systems used for most capacitors are once again based on the resistor system, with the first three colours giving the value in picofarads (trillionth of a farad). In the example of Fig.4, the broad red band at the top of the component is in fact bands one and two of the colour code, and gives '22' as the first two digits of the value. The green band is the multiplier, and indicates that the basic value should be multiplied by 100,000, which gives a total value of 2,200,000 picofarads (2,200,000pF). Dividing this by a million produces a more manageable value of 2.2 microfarads (2.2µF or 2µ2).

The fourth band gives the tolerance in more or less standard resistor-code fashion. However, black can be used for a 20 per cent tolerance rating, and white is sometimes used instead of silver to indicate a tolerance of plus or minus ten per cent, as in this example. The fifth band, if present, is most likely to indicate the maximum operating voltage. It is usually either red (250V) or yellow (400V). The Internet should be able to help in the unlikely event that you obtain a ceramic capacitor having one of the obsolete colour codes. It should be borne in mind that any capacitor of this type is almost certain to be very old stock, and less than ideal for use in a new project.

AUDIO OUT

By Jake Rothman

Test-bench Amplifier debate

John Ellis offers some design comments for improving our recent Test-bench Amplifier

Dear editor, Jake Rothman's *Test-bench Amplifier* in the Nov 2014 issue of *EPE* and the text describing it have several points which reduce its usefulness.

Differential pair tail current

At first sight, it seems there are several basic errors in the circuit. Problem number one, if the input differential pair is operated at 1.6mA tail current, each side should be designed to pass half (800µA). If the second stage (TR4) operates at 6mA, the voltage drop across its emitter resistor (68Ω) is 400mV, making the base voltage around 1 to 1.1V from the local positive rail. The base resistor should therefore be about 1.5kΩ, not 820Ω.

Current source references

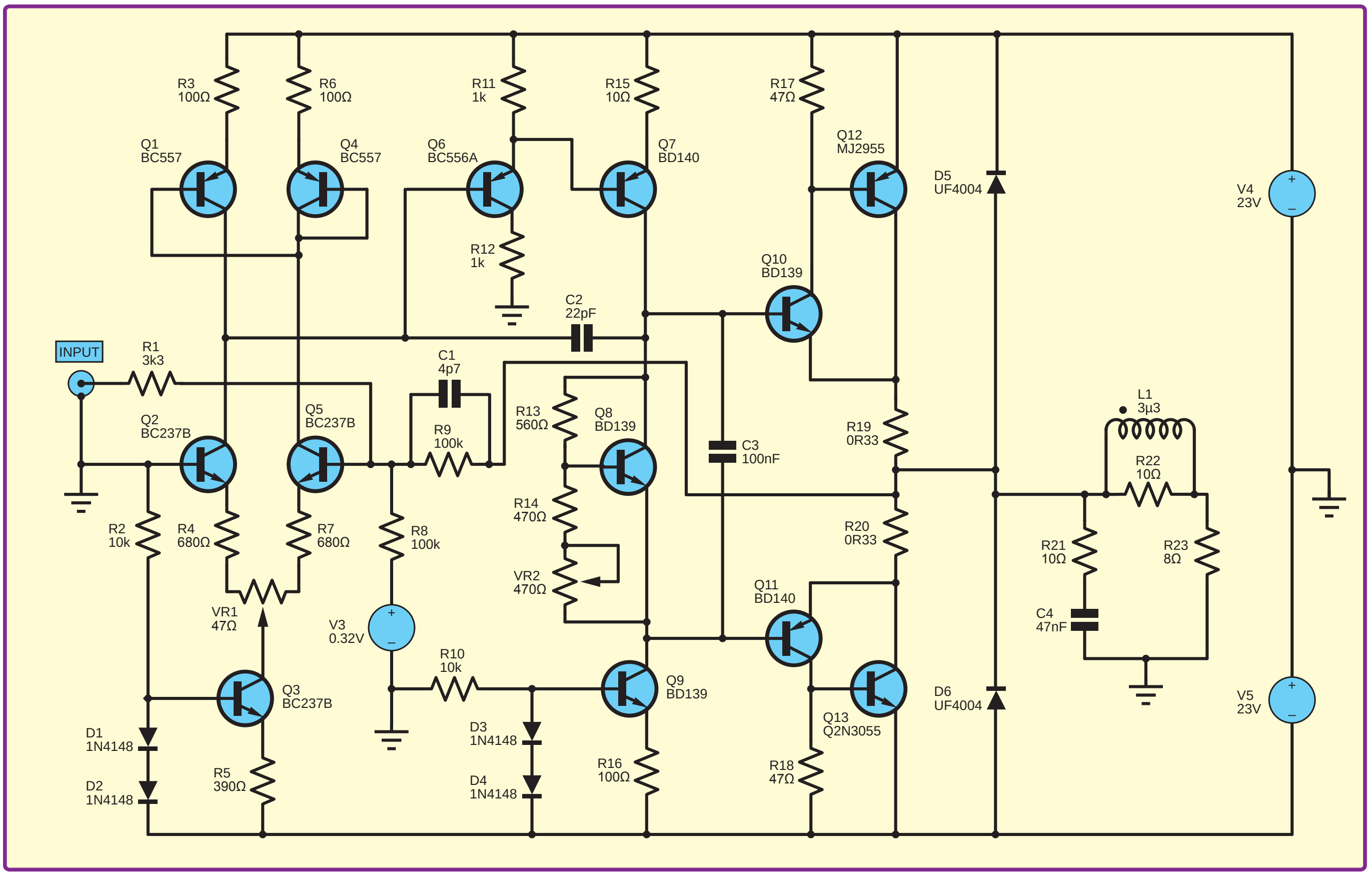
The current sources used for the first and second stages are run from the same Zener diode reference voltage. They have base resistors of 1kΩ each, presumably so that if either were to saturate (and in practice only TR5 might) they would not hog the current from the reference, which would upset the other stage too. Another difficulty is that they only serve to limit the frequency response of the transistor stages by an RC combination with the collector-base capacitance current. In many designs, a simple diode pair can be used as a reference – why not use two separate bias circuits and eliminate the 1kΩ resistors?

Output stage resistors

The output stage used to be widely used, but surely there are errors here too. The low value resistors appear to serve no useful purpose in limiting the quiescent current, and should either have the feedback resistors (62Ω) connected to the junction between one and its corresponding output transistor collector, or should be wired into the emitter of the appropriate output transistors.

Stabilising capacitor

The stabilising capacitor of 470pF is rather large. Given that the input stage operates at 800µA (or would if the bias resistors were right) the input stage gain has a transconductance of 15mA/V, equivalent to 60Ω. The cut-off



A better test-bench amplifier or POPAMP (power op amp), based on the 'Blameless' architecture of Doug Self. This simulates at 0.027% THD at 20kHz and does not suffer TIMD for 20kHz square wave input. V3 represents an offset adjustment and can be made from a 1.25V bandgap circuit to ground with a 10kΩ potentiometer to R8. If positive and negative inputs are needed, a similar offset is needed on the non-inverting input.

frequency (unity gain) would therefore approach 5MHz, which is fine except that the slew rate is set by the input stage current to 1.7MV/s. Add in the 150pF capacitor too and the slew rate becomes even worse – about 1.25MV/s – which means that at 20kHz it will be slew-rate limited. The simulated distortion figure for 15W output is 0.2%, which is not even up to the Hi-Fi standard that the author seems to decry. Why not use a conventional complementary feedback pair, which allows the frequency compensation capacitor to be smaller, improving the frequency response and reducing distortion?

Hum and EMI

Regarding some comments that have been made, we should correct some misapprehensions. The author seems to think that Hi-Fi amps emit excessive hum and EMI. Hardly. Such amplifiers would be 'low-fi' at best, and interference generators at worst. No real Hi-Fi amplifier I have ever built or heard gives out anything more than barely audible hum (if that), nor EMI. Most real Hi-Fi amps also specify high quality transformers. Again, thumps and squeaks, which may occur on turn-on are due to poor design, or architectures.

Feedback

The complementary feedback pair is capable of delivering low crossover distortion, but it is wrong to assume that this is 'for free'. At high frequencies, when one of the output stages turns on at the point of crossing over from the other half, the driver transistors usually experience a current spike to push the power transistors on hard. This momentary increase in current requires a minimal resistance in the driver emitter circuit, which in the existing circuit is not possible without clipping due to the high value of the emitter resistor (82Ω). This is the main reason for suggesting using a standard CFP output. A standard emitter follower, however, can keep the drivers in Class A and give lower distortion.

It is also incorrect to state that the shunt feedback arrangement will give lower distortion than in a standard non-inverting configuration. If hard slew-rate distortion is to be avoided, emitter degeneration resistors will be required, even in the inverting configuration. Especially in the original circuit, due to the slow slew rate as mentioned, negative feedback will be late in arriving for high frequency signals – say 20kHz – and the input stage will be hard pressed to respond.

The 'golden rule' in amplifier design is to ensure that the input stages cannot overload under normal conditions, which essentially means ensuring that the input stage can handle twice the normal peak input voltage without overloading. Then, the gain can be higher in subsequent stages to keep overall distortion low.

Blameless architecture

A better approach is to use the 'blameless' architecture presented by Doug Self in *Electronics World* (see February 1994 issue). I have included an adaptation with some differences in the diagram below. The main features are:

1) Emitter degen resistors are used
2) Darlington voltage amplifier stage is used to restore gain
3) Current mirror is used in the input stage, which provides a doubling in gain compared with the standard design and a higher dynamic impedance compared to a resistor which would otherwise have been required
4) Standard CFP (complementary feedback pair) output is used as this gives good thermal stability without needing to mount the bias regulator on the output heatsink (but this method works well with standard emitter followers if the drivers are also on the same heatsink)
5) Output transistor base resistors are much lower to provide a quicker base turn off drive. Compared with the original blameless circuit, the emitter degen resistors are bigger to suit the lower current operation in the input stage, and the CFP rather than straight emitter followers, and separate reference voltages are used for the input and VAS (voltage-amplifier stage) current sources.

A final few points: first, the original circuit architecture is prone to switch-on surges as it is not symmetrical. However, the proposed current mirror goes some way to ensuring that the circuit turns on at low voltages correctly, since the currents in the input stage will balance over a wide range. A simulation of a slow turn-on shows that only a 300mV excursion occurs.

Second, the asymmetry also gives rise to wider variations in output voltage with temperature. Some means of providing an input bias current which is temperature and gain compensated is needed to minimise output voltage zero drift, but being unbalanced there will be some offset due to changes in the base current of the voltage amplifier stage input transistor with temperature. It is tempting to use FETs in the input stage to eliminate base bias currents.

Third, it might seem a nice idea to use a PTC within the feedback loop, but if the feedback loop is disconnected and there is no input offset provided, as required for zero drift cancellation in any case, the output could float to any voltage such that there might be a large 'thump' when the PTC switches back on. It would be better to incorporate a standard protection circuit for the power transistors.

Fourth, if the input stages were to be replaced by an IC there is no guarantee that the slew rate can be adequately controlled as the degeneration resistors may or may not be present, or if present at the optimum value needed. For this reason, discrete power amplifiers are still generally better than IC designs. It would, however, be possible to hook up an IC using level shifters while still using local Miller feedback from the pre-driver stage. More commonly, a higher voltage is used for the input stages than the output rails.

And my last point is that generally speaking the classical Miller compensation method is my least favourite means of stabilising an amplifier. However, I do agree that it is the right approach for a test-bench amplifier, which is indeed effectively a power op-amp, where feedback networks may be capacitive or inductive as perhaps might be called on for analogue computing. The one main advantage over other methods is that it is unconditionally stable (in principle) and reactive loads will (generally) not be a big problem.

Jake Rothman replies

First, I thank Dr John Ellis for taking an interest in the design and taking its theoretical analysis to a higher level. On this basis I agree with his improvements and appreciate his scientific rigor, but pragmatic concerns about ruggedness and simplification of the circuit, have involved some compromise compared to the optimal Hi-Fi amp he would design. I'll address the points in order.

Input stage

John is right about the dimensioning of the input stage resistor values and also about the slewing problem.

TR1's collector resistor value was tweaked from optimum to enable the DC offset at the output to be easily adjusted to zero. The current through TR1 was measured at about twice that in TR2. Since feedback was applied across an output capacitor, the overall DC feedback resistor had to be very high at 330kΩ and this made the circuit more sensitive to base bias currents. This was an engineering compromise of cancelling two errors. In view of the small increase in second-harmonic distortion caused by this current

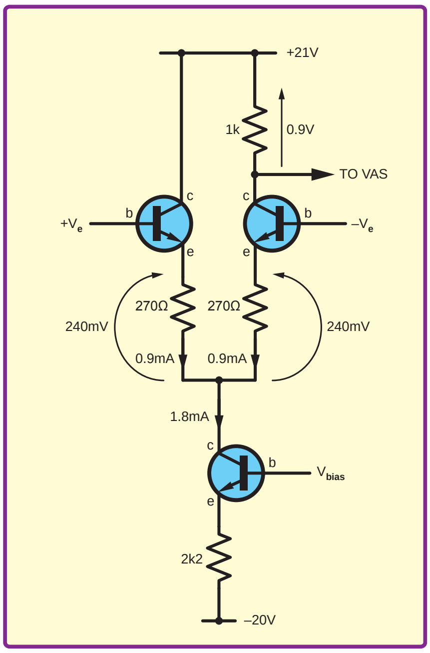


Fig.1. Revised input stage with changes including emitter degeneration

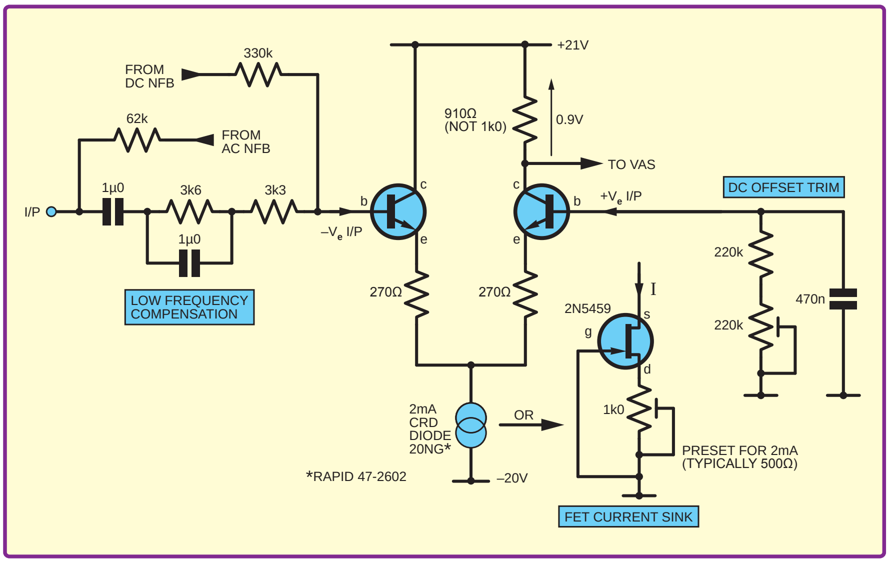


Fig.3. Isolating current sources – a separate current sink will stop interaction if the VAS current sink saturates. A current regulator diode can be used or its discrete FET equivalent. Note the revised low-frequency compensation and extra resistor on DC offset trimmer

imbalance compared to the much greater distortion in class AB output stages, it seemed a reasonable compromise at the time. All the values were then re-tweaked to get the currents equal. Fig.1 shows the new values of currents, voltages and resistors. In the end, I increased the current a bit, increased the load resistor to 1kΩ and added 270Ω degeneration resistors.

Note that the offset null pot needs a 220kΩ resistor in series with it (see Fig.3). Alternatively, a fixed 330kΩ resistor could be used (the same as the DC feedback resistor) since any offset is blocked by the output capacitor.

Slewing

This was an oversight that did affect the sound a bit, not to my ears where I can only hear up to 13kHz. But my wife and son who can hear up to 19kHz noticed. The amplifier did indeed have slew-induced distortion at 20kHz, as shown in Fig.2. My fault, I never tested it at a high level at 20kHz. The fact it was still there when the output was unloaded

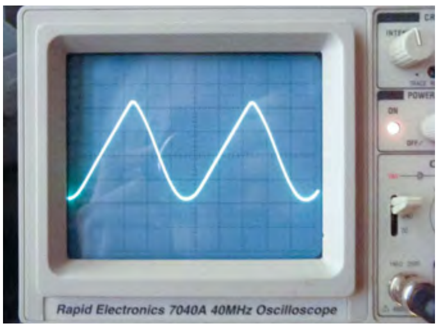


Fig.2. Oscillogram showing the characteristic ramp profile of slewing-induced distortion at 20kHz and 12V pk-pk. Note the asymmetry, possibly due to the unequal currents in the LTP (long-tailed pair)

showed it was an input stage problem. Interestingly, with loud percussive jazz I could hear intermodulation products going down into the audible region when flipping between modified and unmodified amplifiers.

Luckily, there's a tweak costing a penny to fix (as suggested by John) – reducing the gain of the input stage by adding degenerating resistors to the emitters of the long-tailed pair. The standard trick here is to reduce the gain of the input stage by 10× and then reduce the compensation capacitor by 10× (see Fig.4) to ensure the same closed-loop stability. This then reduces the loading at high frequencies by 10×. With the stage running at the same current the slewing problem is moved up by a decade, out of harm's way.

The calculations are described in detail in Bob Cordell's book, *Designing Audio Power Amplifiers*. Magazines such as *EPE* are practical, so I tend to avoid filling pages with mathematics. However there's nothing so practical as a good equation – where would we be without Ohm's law? In this case, the internal emitter resistance r_e is set by I_e = 0.9mA: 26/0.9=29Ω (remember r_e =26Ω at 1mA and gets proportionally lower as the collector current increases). To get the gain down by 10× we need to increase the total emitter resistance to 290Ω, which would mean adding 260Ω. 270Ω is the nearest E12 value.

Constant current generators

I agree, separate current source bias voltages are a good idea in preventing interaction when clipping occurs. However, once clipping has occurred all battles against distortion are lost anyway. It does mean the clipping is cleaner if separate bias sources are used. I chose the design route presented because there was only one bias source on the legacy Maplin PCB used. Again, it was a pragmatic compromise. Boards were built using separate current regulator diodes (CRDs) and also FETs to replace the biased bipolar current sinks that reduced the component count and avoided the interaction problem (Fig.3) but this was at the expense of using unusual parts. The 1kΩ resistors were used to prevent DC blow-up paths

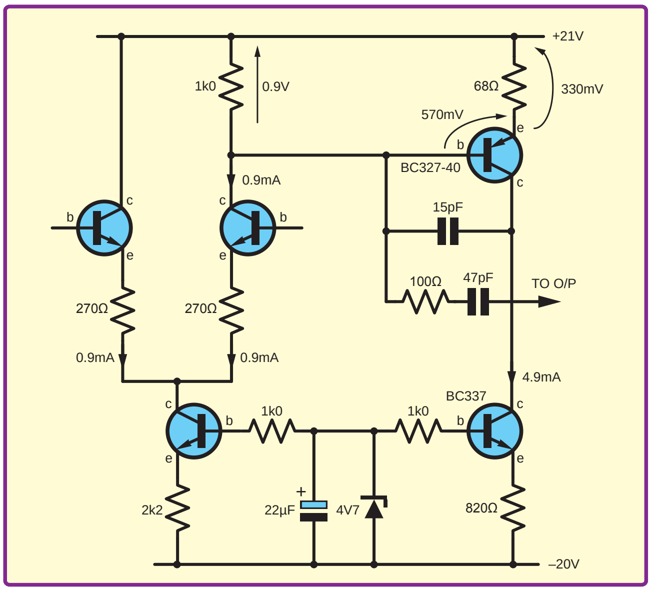


Fig.4. LTP load resistor, VAS and its associated current source. Note HF compensation capacitors are now reduced by 10×

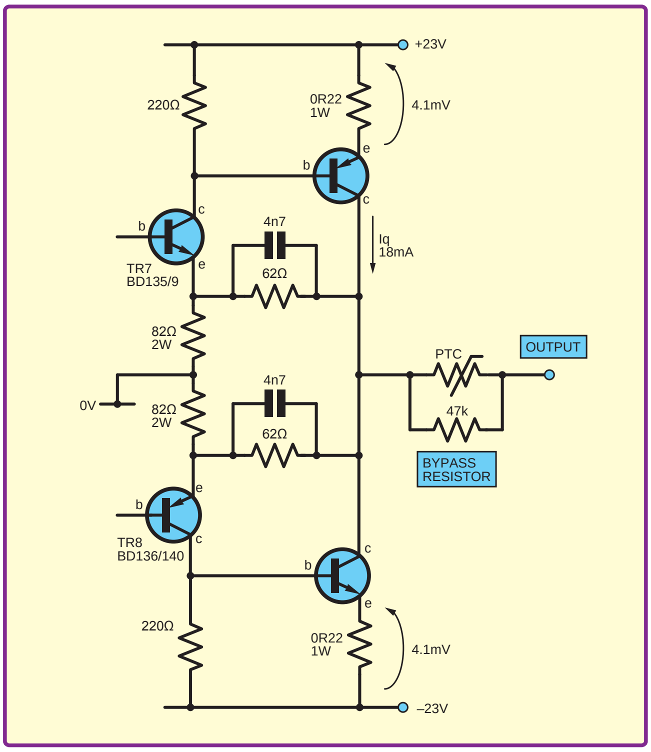


Fig.5. More stable position for output stage emitter resistors. Note bypass resistor across the PTC device

causing a domino effect in the event of component failure or probe slips. I tried bypassing the 1kΩ resistor with a capacitor, but that made the HF output lower – more investigation needed!

I noticed the diagram (Fig.1, *Audio Out*, *EPE*, November 2014) showed an incorrect 820Ω resistor for the LTP (long-tailed pair) current sink. Initially, I used a 2.4kΩ value to set the current to 1.6mA. For the re-jigging, I used a 2.2kΩ resistor, which gave 1.8mA. I also noticed the current in the VAS was 4.9mA not 6mA. (Most of the system gain lies in the VAS – Voltage Amplifier Stage – which lies between the LTP and the output driver stage.) There was now 0.9V across the load resistor which meant its value needed to be 1kΩ to give a current of 0.9mA through one side of the LTP to ensure the currents balance – see Fig.4. Of course, this depends on the V_{be} of the VAS transistor. In this case it was 0.57V.

Compensation

I decided to compromise high-frequency distortion for ruggedness. I have done this in several areas of the design in view of its use as a *Test-bench Amplifier*, rather than a top-notch Hi-Fi amplifier. Indeed, the amp is over compensated to reduce the possibility of oscillation and its possible destruction in unusual test situations. The max output at 20kHz is limited to 16.5W into 8Ω compared to 19.5W at 1kHz.

Adding the degeneration reduces the AC feedback factor by 10×, which will increase the low-frequency distortion by a corresponding amount. This did not show up on listening. The reduced open-loop gain now means the LF compensation is excessive and the network will need changing to maintain the bass response. The values now need to be 1µF and 3.6kΩ, as shown in Fig.3. This just shows the juggling of component values necessary in any design to achieve a reliable compromise. Further optimisation of component values is always possible.

Emitter resistors

The 0.1Ω resistors are so low they do little to stabilise I_q, but they enable it to be measured. I do think thermal stability will be enhanced if they are put in the positions John suggests in the emitters of the output transistors. This was done in the original Maplin design and recommended by the Mullard designer Tharma in his book *Transistor Audio Amplifiers*. The resistors ended up in this design in the position shown by accident, since I was using them as current sensors for testing a traditional short-circuit protection circuit. I agree it is sub-optimal and should be changed.

If a conventional CFP (complementary feedback pair) output stage was used, then the output pairs effectively become complementary-Darlington transistors and the resistors can then go back on the collectors. However, the conventional output topology was not used because of its lack of current limiting. Again, another compromise of ruggedness over *finesse*. I decided to move the resistors to the emitters as suggested. I used 0.22Ω metal-film fusible resistors since they are non-inductive and provide protection in the event of the output transistors going short circuit. Note that the optimum I_q is now around 18mA, or 4.2mV across each 0.22Ω resistor, as shown in Fig.5.

Noisy Hi-Fi

Most commercial Hi-Fi amps emit audible noise when the speakers are close to the ears – in my experience. By Hi-Fi amps I mean standard consumer types such as Sansui, Akai, Trio – not expensive designs. I used to employ an expensive Cyrus II on the test bench, which was quiet, but my wife commandeered it! I also blew it up twice and took out some speakers with it.

CFP output

The CFP output stage with gain does have inferior distortion compared to

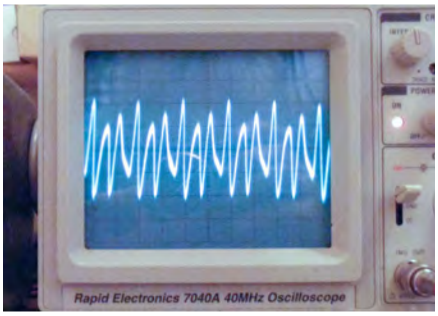


Fig.6. Distortion residuals while setting Iq preset using an 8Ω load 3V pk-pk output (140mW) @ 1kHz. The effect of too little current, spiky crossover

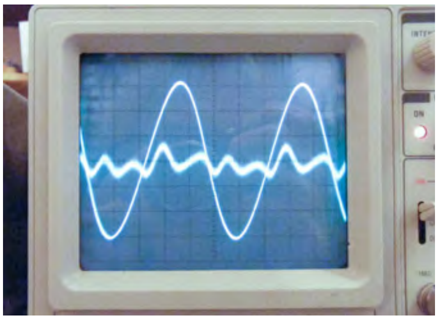


Fig.7. Optimum (18mA) Iq giving 0.033% THD+N at 140mW

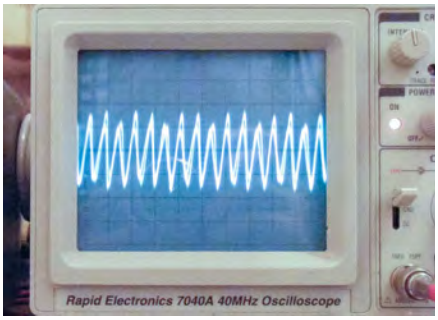


Fig.8. Too much current, resulting in more distortion and possibly thermal runaway

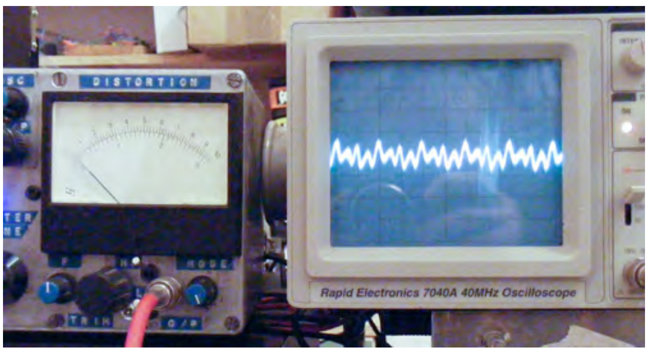


Fig.9. Optimum residual 3mV with output at 25V (10W) at 1kHz corresponding to 0.012% THD+N

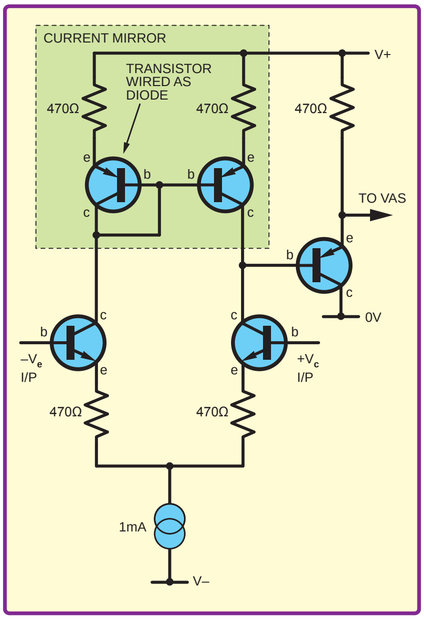


Fig.10. Self/Cordell input circuit – note the use of current mirror and VAS buffer

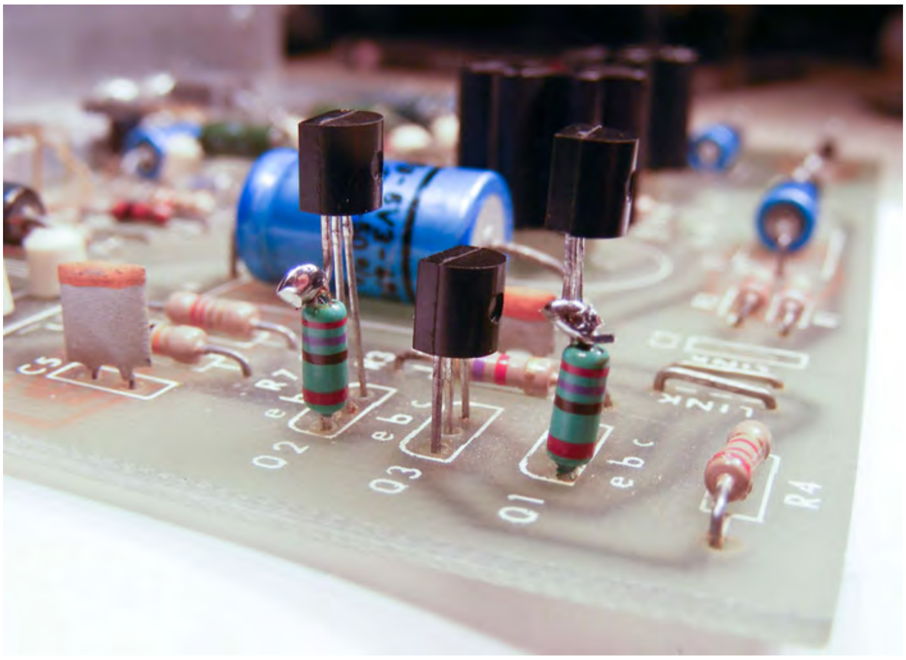


Fig.11. Adding emitter degeneration resistors to the long-tailed pair transistors – a 'penny tweak' that reduces distortion

the emitter follower, which has a gain slightly below one. If I had used this, I would have had to use voltage doublers or extra transformer windings to get the lower noise of a decoupled driver stage for the same output. I would have also needed to incorporate a standard current-limiter circuit.

Distortion

The switching distortion mechanism outlined due to a lack of drive (because of the 82Ω resistors) manifests itself as increased crossover distortion seen in the residual. But it is also this lack of drive that helps provide the required protection. I show the distortion residuals in Fig.6, 7, 8 and 9. The residual distortion after filtering out the fundamental is about 3mV pk-pk at 25V pk-pk 10W output, equating to around 0.012% THD and noise. This doubles at 10kHz due to decreasing feedback margin. At 100Hz it's the same, showing the feedback around the output capacitor is doing it's job.

Doug Self's amplifier

I really admire Douglas Self's 'blameless' architecture and with justification it has become the *de facto* standard optimal Hi-Fi topology. But it is too complicated for this *Test-bench Amplifier* and I've seen so many versions of it with minor component value changes, I'm sure many readers are getting bored. Every serious audio designer I know has his books. If factory assembly is used it makes economic sense to use lots of extra components to get distortion a little lower, but for the home constructor/experimenter, fewer parts mean generally mean fewer construction errors.

If readers want to add a current mirror to enforce equal currents (as in Fig.10) it will be a good experiment to do. I would like to check it with unmatched transistors. It also has the benefit of removing the 1kΩ collector loading resistor, replacing it with a dynamic load of around 40kΩ. Also, since the mirror works in push-pull, the transconductance is doubled again. However, the loading of the VAS now becomes much more significant, so this needs to be buffered to get the full benefit and we now arrive at the Self/Cordell input circuit shown in Fig.10.

Thumps and symmetry

I disagree that symmetrical circuit design is important in preventing thumps, some of the lowest thump circuits I've seen even use single power rails. The designs popularised by John Linsley-Hood and Rod Elliot of ESP, all used skilful ramping of bias voltages and currents. As for using a simulator to test for transient power up/down anomalies, I have grave doubts about this approach, since a lot of variables will be undefined and the system is essentially chaotic during these periods.

Also, I suspect many an audio designer's love of symmetry is often based on human aesthetic judgement. The circuit diagram looks prettier, therefore it must be better. It's true, symmetry does give a more stable DC performance, but we are designing for audio, not instrumentation. The standard LTP itself is implicated in causing thumps, because it tends to 'snap-on' when powered up within a negative-feedback loop. I look forward to trying a current mirror to see if it is capable of reducing this problem. Single-ended inputs and the complementary LTP can be biased up smoothly. Some engineers regard the cancellation of the even-order distortion harmonics in symmetrical circuits as a bad thing, since they are 'enhancing' unlike the dissonant odd-order harmonics that remain.

PTC device protection

I've found the use of a PTC protection device (here Raytheon's 'Polyswitch') does not cause a thump when activated and deactivated. It does not disconnect the feedback since it is not a proper switch, it just goes to a high resistance state (approximately 90kΩ) so no massive offset is generated in this circuit. However, if a normal fuse is used, a bypass resistor (Fig.5) needs to be wired across it to maintain the loop. It might be a good idea to use one with a PTC just in case, since the 'off' resistance may vary a lot between devices (no I haven't checked yet). The slow thermal switching time of a PTC just slowly ramps any increased offset and is inaudible.

Discrete design

I'm glad John agrees discrete design gives better audio performance than chips, because everything is under the control of the designer, as well as being much more interesting. I do appreciate his suggestions, which have improved the design. Audio engineering is applied science, and the scientific principles of experimentation, measurement, replication of results and peer review are essential for progress.

Constant current sources

REGULAR *EPE Chat Zone* contributor ***afferrari*** posted a question about constant current sources.

After reading about constant current sources, I defined the four cases shown in the PDF [Fig.1]. Current should go up to 250mA. Implementing B and C with the attached circuit [Fig.2] (and its counterpart) is straightforward, but I am not sure how to proceed for A and D. I cannot have R_{sense} referred to ground anymore. How could I proceed without fancy ICs?

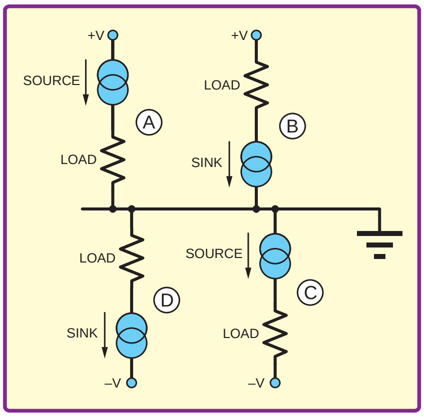


Fig.1. Current source cases from ***afferrari's*** Chat Zone *post (see question text)*

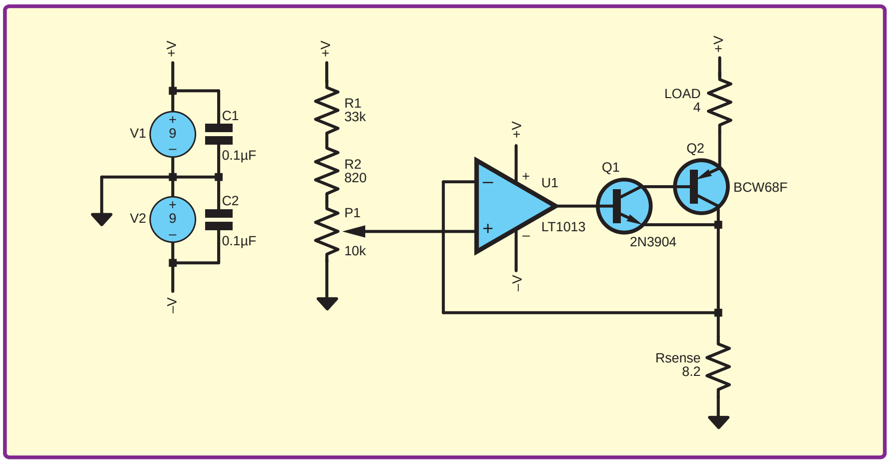


Fig.2. Schematic from **afferrari's** Chat Zone *post (see question text)*

The schematic of a possible solution was provided by another regular *Chat Zone* contributor, ***bowden_p***, followed by some discussion and another circuit from ***afferrari***, who seemed happy that the issue was resolved. This is a good example of the *Chat Zone* in action – readers who have not seen the discussion are recommended to take a look and hopefully join in too (**chatzones.co.uk**).

Although the specific point of this question was answered, current sources are an interesting topic, and one that sometimes causes confusion, perhaps because people are simply more familiar with voltage sources. In this *Circuit Surgery* we will look at the concept of current sources and a basic current source circuit. Next month, we will discuss op-amp-based current sources, like the circuits posted by ***afferrari*** and ***bowden_p*** in the *Chat Zone* thread that prompted this article.

Constant current sources are very important in electronic circuit design; they are to be found inside almost every analogue IC (such as operational amplifiers) where they have widespread general use for biasing and active loads. More specific current source uses include LED drivers, battery chargers and ramp generators (a constant current applied to a capacitor gives a linearly increasing voltage).

Ideal and real sources

Before we look in more detail at current sources it will help to look at voltage sources, so we can see some basic concepts in a more familiar context and then compare the two types of source (see Fig.3 for their symbols).

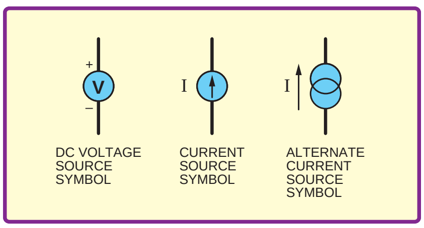


Fig.3. Voltage and current source symbols. Unless otherwise stated these represent ideal sources

An ideal constant voltage source can output any current without the voltage changing, or cutting out. Ideal sources are mathematical models and do not exist in the real world (this applies to both current and voltage sources); however, they can be very useful in simplified calculations and simulations. For example, when analysing circuits we often assume that the power supply is an ideal voltage source. In practice, of course, this is not true – a battery's voltage will drop if you connect a heavy load (low resistance), a bench power supply may cut out or limit the current when overloaded.

Batteries, the mains and bench power supplies approximate ideal voltages sources to some extent; however, commonplace constant current sources are less obvious. Solar cells provide a more or less constant current output for a given light intensity and temperature, at least while the output voltage is relatively low, so they may be the best example.

To look at current and voltage sources in more depth we can consider what happens when we connect a load resistor across the source – using Ohm's law – the basic relationship between current, voltage and resistance. This is $V = I \times R$; so, to find the voltage across a resistor (V in volts) multiply the current through it (I in amps) by the resistance (R in ohms). If we connect a voltage source

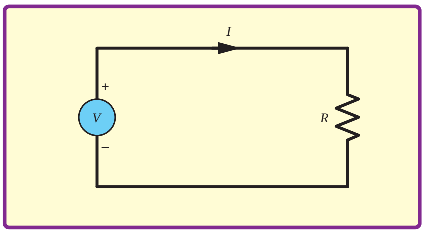


Fig.4. Ohm's Law: V = IR

to a resistor, as shown in Fig.4, it must produce a current given by $I = V/R$. If we connect a resistor with a very low value across the voltage source it must produce a very high current in order to satisfy Ohm's law for the resistor.

For example, a 9V battery with a 0.1Ω resistor across should produce a current of 90A (9/0.1 = 90), but a small battery like a PP3 is simply not up to the task. So what happens if the voltage source is not capable of producing the required current? Is Ohm's Law broken? No; the voltage source pushes out as much current as it can and its output voltage reduces to the point given by Ohm's Law for the load resistance and current involved.

Internal resistance

We can understand what is going on using the concept of internal resistance. Real voltage sources can be thought of as comprising an ideal voltage source (with voltage V_S) and a resistor (R_{int}), as shown in Fig.5.

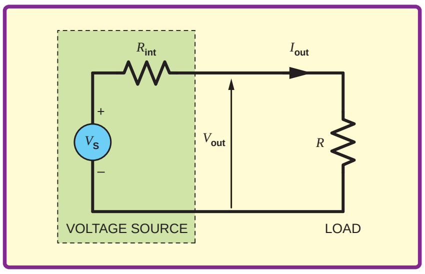


Fig.5. Internal resistance of a real voltage source

So when we connect a resistor to a real voltage source the current (I_{out}) flows through both resistors, with the voltage likewise dropped across both components.

The voltage at the terminals of the real source (V_{out}) therefore becomes smaller and smaller as R is reduced and a greater proportion of V_S is dropped across R_{int}. In practice, we cannot physically separate the voltage source and internal resistance. A battery's internal resistance is noticeable as a drop in voltage as we decrease load resistance, and as its internal resistance increases over its service life the resistance tends to get worse as the battery is discharged.

The maximum output voltage from a real voltage source occurs when the output is open circuit, so we may state the *open-circuit output voltage* (equal to V_S in Fig.5). The internal resistance limits the maximum output current to V_S/R_{int}, at which point V_{out} is zero. The maximum current is delivered into a short circuit (R = 0 in Fig.5) so we have the *short-circuit output current*. For an ideal voltage source the short-circuit output current is infinite.

The effective internal resistance of bench power supplies, regulator chips, and other circuits can be made very small using feedback control circuits. These monitor V_{out} and effectively change R_{int} or V_S so that V_{out} remains constant, compensating for the changing drop across R_{int}. This cannot go on forever of course, because at some point we reach the maximum current or voltage rating of one or more of the internal components, or some other limiting factor, at this point the source may cut out or limit the current, which will cause the output voltage to drop.

Bench power supplies, for example, can sometimes deliver the same voltage (with little drooping) until some maximum current at which point they cut out. Here, due to the active control circuitry, the maximum output current may be much less than that implied by the effective internal resistance (V_S/R_{int}). Thus a good bench power supply may be a good approximation of an *ideal* voltage source, but only over a limited range of currents.

Current sources

An *ideal* current source outputs a particular current irrespective of the voltage, but *real* current sources are limited in terms of the range of voltage over which they will operate. Real current sources also have internal resistance (see Fig.6), but it appears in parallel with an ideal current source and the larger it is the better the current source. Compare this with a voltage source where the internal resistance is in series and smaller the internal resistance the more ideal the source is.

If we connect a large resistance across an ideal current source, then by Ohm's Law it must develop a large voltage across the resistor. For example, if we connect a 1MΩ resistor across a 1mA current source it would have to produce 1000V across the resistor to maintain the required current, probably impossible for a current source circuit with a low-voltage supply such as the one posted by ***atferrari***. Just like

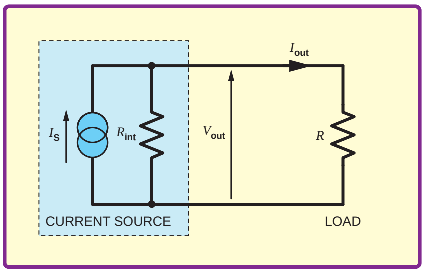


Fig.6. Internal resistance of a real current source

the battery and bench power supply discussed earlier, real current sources have their limitations.

The internal resistance, R_{int}, of a current source, I_S, limits its output voltage, V_{out}, to $I_S R_{int}$ (Fig.6), at which point I_{out} is zero – so this is the open-circuit output voltage. The larger the value of R_{int} the larger the maximum output voltage. For ideal current sources the open-circuit output voltage is infinite and R_{int} is infinite. If we short circuit a current source (R = 0 in Fig.6) the output current is equal to I_S and there is zero voltage across the source.

Real current sources may also have a limited range of voltages over which they can operate (sometimes called 'compliance'), determined by factors other than the internal resistance (in a similar manner to the bench power supply described above).

An ideal voltage source delivers infinite current to a short circuit and an ideal current source produces infinite voltage across an open circuit. Note the opposites here: real voltage sources have an 'easy life' with open circuits and a 'hard time' with short circuits, for current sources it is the other way round.

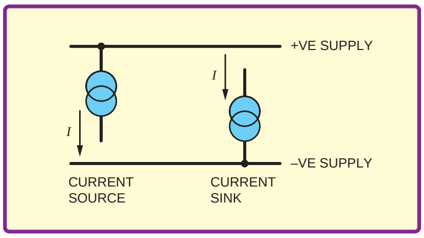


Fig.7. Current sources and sinks

Although the term 'source' can be used for all current sources we sometimes makes the distinction between current sources, from which current flows *out*, and current sinks, to which current flows *in* (see Fig.7). This is also seen in ***atferrari's*** diagram (Fig.1).

Circuit theory

The current source and voltage source in Fig.5 and 6 have a more general use in circuit theory than just representing power sources such as batteries. Thévenin's and Norton's theorems allow any two terminal circuit output to be represented this way.

Thévenin's theorem states that any network of current sources, voltage sources and resistors, which has two output terminals, can be represented as a single ideal voltage source with a single series resistance (like Fig.5). This is the known as the Thévenin equivalent circuit. Norton's theorem is similar except that the equivalent circuit comprises an ideal current source and a parallel resistor (like Fig.6). It is possible to convert between the Thévenin and Norton equivalents. The internal resistor

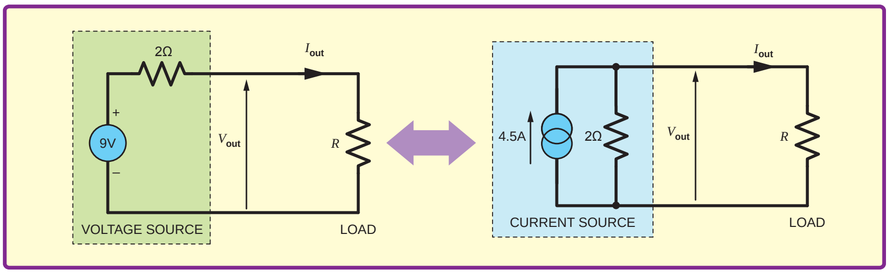


Fig.8. Equivalence between current source and voltage source representations. This shows a reasonable constant voltage source is equivalent to a poor constant current source

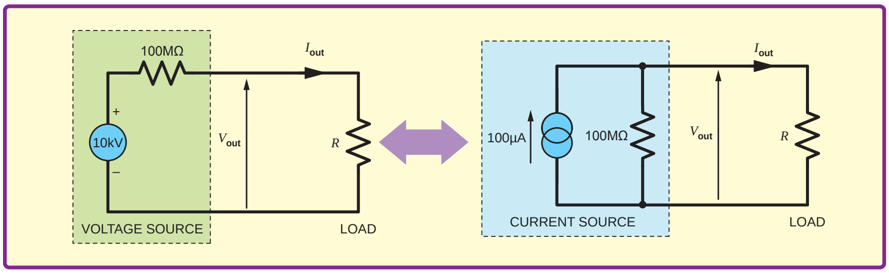


Fig.9. Another source equivalence example. This shows a good constant current source is equivalent to a poor constant voltage source

value R_{int} is unchanged and the source values are calculated using:

$$V_{Th} = I_{No}R_{int}$$
$$I_{No} = V_{Th}/R_{int}$$

in which the subscripts for the sources *Th* and *No* refer to the Thévenin and Norton equivalent circuits respectively. The Thévenin and Norton versions of the source are not equivalent to one another or a circuit they represent in all respects – their internal power dissipation is not equal.

As an example, consider a 9V battery with a 2Ω internal (series) resistance; this is equivalent to a 4.5A current source with a 2Ω internal parallel resistance (see Fig.8). We can check that these are indeed equivalent by calculating what happens when we connect a load resistor across the terminals. If we use 400Ω in both cases we get a current of 22.4mA through the load resistor and a voltage of 8.96V across it, reduce the load to 300Ω and we get 29.8mA and 8.94V.

We see these voltages are very close to the Thévenin equivalent voltage, but a long way from the Norton equivalent current, with the change in load causing a large proportional change in load current (+33%), but changing the load voltage only a little (–0.2%). As we might expect, our battery is behaving as a reasonably good constant voltage source, but as a very poor constant current source.

As another example consider a 10kV voltage source with a 100MΩ internal resistance. This is equivalent to a 100µA current source in series with a 100MΩ internal resistance (see Fig.9). If we connect a 100kΩ resistor across the output we get a current of 99.9µA and a voltage of 9.99V. If we increase the load to 200kΩ the voltage almost doubles to 19.96V, but the current remains almost the same at 99.8µA. This situation represents a good constant current source and poor voltage source. Using a very high voltage source and large resistor like this is not usually a practical way to produce a current source circuit.

Thévenin and Norton equivalent circuits are used in circuit analysis to reduce complex circuits to a simpler representation for the purpose of further analysis. Using one or the other equivalent does not imply a circuit acts anything like an ideal current or voltage source; but if you do find these equivalent circuits the internal resistance value will indicate if the circuit may behave more like one ideal than the other.

Practical current sources

Transistors provide us with a ready means of creating a current source. Bipolar transistors work by using a voltage at the base to control the current flowering from collector to emitter. If we make the voltage on the base of a transistor constant we get a constant collector current, so the collector acts as the output of a current source (see Fig.10a).

You may be thinking that it is the base current which controls a transistor's collector current, but there is a fixed relationship between base-emitter voltage and collector current, so a given base voltage results in particular base current (for a given transistor at a fixed temperature). The reason we often think first in terms of base current is the simple relationship between base current and collector current used when considering amplifier circuits (the transistor's current gain), whereas the base-voltage to collector-current relationship involves a more complicated exponential equation.

We could obtain a fixed voltage for our transistor's base using a potential divider (see Fig.10b). This circuit does not work very well because the current changes with temperature – all semiconductor devices such as diodes and transistors are very temperature sensitive. The current will also change quite a bit if the supply voltage changes due to the exponential base-voltage to collector current relationship.

A better approach to obtaining a fixed base voltage uses another transistor connected as a diode. If we short the collector and base of a transistor together we are left with the base-emitter PN junction – the

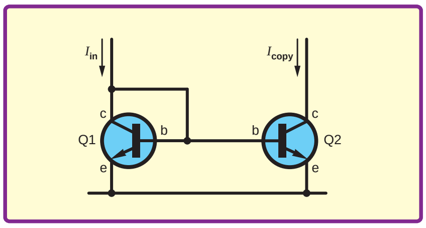


Fig.12. The current mirror

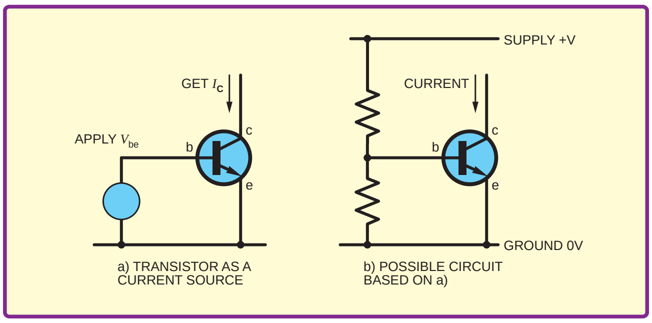


Fig.10. The transistor as a current source

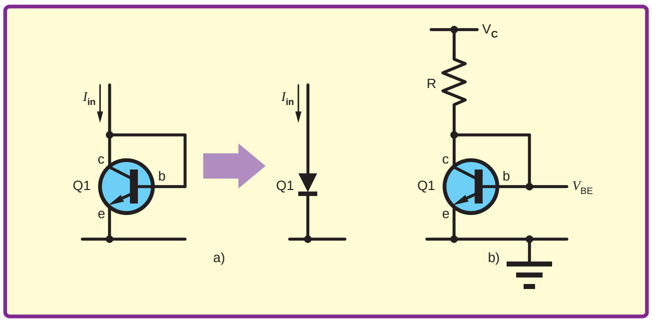


Fig.11. Diode-connected transistor

transistor acts as a diode (Fig.11a). We can bias this diode to carry any reasonable forward current by connecting a resistor from the supply as shown in Fig.11b. The forward voltage drop of the 'diode' will be the V_{BE} of the transistor with a collector current as set by the resistor.

If we wire this diode-connected transistor to another transistor – emitter to emitter and base to base – they will both have the same V_{BE} (see Fig.12). If the transistors are of the same type and at the same temperature the equal V_{BE} implies equal collector currents. Thus, whatever current is set for the first transistor will also flow in the second ($I_{copy} = I_{in}$ in Fig.12). This circuit is known as a 'current mirror'. Current mirrors take a current and produce a copy of it. A similar circuit can be built using MOSFETs.

To get a constant current source (rather than a current mirror) we supply a fixed input current through the diode-connected transistor, as in Fig.11b. This leads to the current source circuit shown in Fig.13. To calculate the value of R we assume that the base-emitter voltage of the transistor V_{BE} is fixed, and in the range 0.6-0.7V, the output current is then

$$I_{out} = (V_{CC} - V_{BE}) / R$$

If we have more details of the transistor, such as the V_{BE} *vs* I_C curve we can set this more accurately. This current will only be constant if the supply is constant, if this is not the case more sophisticated 'self biased' current source circuits can be used.

If the two transistors in Fig.12 and 13 are identical, then because their base-emitter voltages are the same they will have the same collector current (except for a small error due to the base current of Q_1). What exactly do we mean by identical transistors? Transistors are identical or more realistically very similar (known as matched transistors), if they have the same parameters (gain etc.) and the same physical conditions (temperature etc.). This can be achieved by making them the same size and shape and close together on the same piece of silicon.

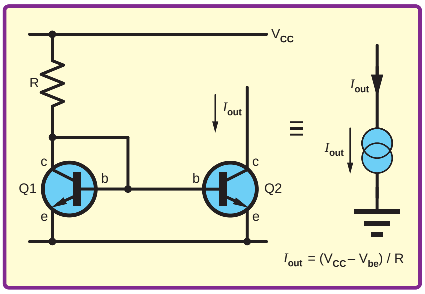


Fig.13. Constant current source (sink) based on a current mirror – Q1 and Q2 are matched transistors

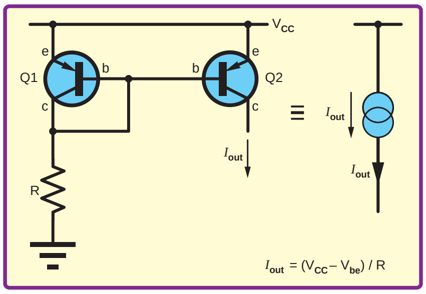


Fig.14. Constant current source

Chip designers know various tricks to make sure transistors are closely matched. Discrete matched transistor pairs are also available.

Limitations

Like all real current sources the circuit in Fig.13 will have internal resistance, as shown in Fig.6. This will be equal to the transistors output resistance, which is caused by base width modulation, also known as the Early effect. The Early effect causes variation of collector current with changing collector-emitter voltage (with fixed base-emitter voltage). An ideal transistor's collector current would depend only on base-emitter voltage and not be affected by collector-emitter voltage. Various other current mirror/current source circuits are available which significantly improve the internal resistance compared with the circuit in Fig.13.

The circuit in Fig.13 has a limited range of voltages (at the collector of the output transistor) over which it will work. The circuit will stop working once the output collector voltage drops below about +0.2 to +0.3V (the transistor's saturation voltage, V_{CEsat}). There is also, of course, a maximum voltage beyond which it will not work due to stress or breakdown of the device.

In ***atferrari's*** post he refers to both current 'sinks' and 'sources', the circuit in Fig.13 is a sink in these terms; the equivalent source circuit is shown in Fig.14. Fig.13 corresponds with cases B and D in Fig.1 and Fig.14 corresponds with cases A and C. These circuits do not have a sense resistor, so at this point in the discussion the R_{sense} issue does not arise.

In this article we have introduced the concept of current sources and looked at some of the fundamental circuit theory relating to them. We have also looked at a simple current source circuit. Circuits like the ones discussed here are often used for bias and loading in analogue ICs because they are relatively small (few transistors). For higher performance a different approach can be used in which a feedback circuit senses and regulates the current. This is the type of circuit ***atferrari*** was discussing (Fig.2) and we will look at this next month.

PIC n' Mix

Mike Hibbett

Our periodic column for PIC programming enlightenment

Hardware hackathons

WE took some time out back in November last year to take part in an event that is becoming increasingly popular in the 'Maker' community – a weekend hardware hackathon.

'Hackathons' are events where a group of people get together for a weekend and spend the entire weekend – typically without breaking to sleep – working in small teams on creative projects. Normally, the idea is to start completely from scratch and aim to have a working project or prototype at the end of the weekend. Sometimes there are prizes; more often, people take part just for the fun of it.

Software hackathons have been around for a couple of decades, but the last few years have seen the rise in popularity of the hardware hackathon, born from the 'Maker' movement of the last decade. Dublin is a major European technology hub, so it's no surprise to discover that there were at least four hardware hackathons in Dublin alone during 2014; *EPE* popped along to one in November to find out what it's all about.

Fig.2. Part of the 'Drum Pants' team hard at work!

SciHackDay

The 'SciHackDay' concept started in the US, and was picked up in Dublin four years ago. Despite the name, it's actually a two-day event, starting mid-morning on a Saturday, running through to mid-afternoon on Sunday. Afterwards, there is a 'show and tell', followed by a light-hearted award ceremony, where laser-cut wooden medals are given out as prizes. With a range of award categories that would have made the Pony Club proud, virtually everyone gets a medal.

Unlike some of the more commercially oriented hack events, Dublin's SciHackDay event is free to attend and with the help of sponsors and volunteers they even managed to provide free food and refreshments for the entire weekend.

This event was held in a suite of offices within the innovation centre of one of Ireland's larger universities, UCD. The university loaned us a dozen 3D printers for the event, but they didn't get a huge amount of use – people typically brought everything they needed. The venue provided desks, chairs, lighting and electricity, but little else.

Deciding exactly what to bring to these events is quite a challenge. Soldering iron, solder, cutters, hot melt glue gun, the electronics you intend to use are obvious choices – but these events can take you down unusual routes, so it's advisable to bring along as wide a range of tools as possible, and a good stock of components. Thankfully, these are very inclusive events and although we formed teams to work on separate projects, it was a highly non-competitive event and people were happy to share their tools, components and even programming skills.

Fig.1. Early morning start

At the event

Saturday morning started quietly enough, with people ambling in during the morning, eating apple lattices and discussing their project ideas with the organisers. About 70 people came along, a mix of scientists, engineers, hobbyists and some curious folk who stumbled upon us by accident, and who decided to stay. Around 11am we were herded into one conference room where each project team had a few minutes to pitch their idea. It's all very, very informal – the complete opposite of 'Dragon's Den'. (Bear in mind, should you be looking to take part in a hardware hackathon, that many events are *exactly like Dragon's Den*. However, SciHackDays and Maker events in general tend to be the exact opposite.)

Drum Solo Pants

The emphasis for the day was projects with a science basis, but the event was open to all ideas covering software to hardware. Our project was definitely in the 'other' camp.

We had to develop a pair of trousers fitted with sensors, which would trigger drum sounds on a remote sound system – 'Drum Solo Pants'. The idea was to take my 'Open source drum kit' concept and try to recreate it on an Arduino and Raspberry Pi platform. The question being, can we do this just as well, but much more quickly on what is supposed to be a pair of easy-to-use software development platforms. Rather than have the drum pad sensors wired directly into a dsPIC microcontroller, the sensors would couple into an Arduino platform, and then, via, a wireless interface, to a Raspberry Pi, which would host the sound samples.

With the Arduino and Pi being established development platforms, the main issue revolved around the wireless interface. Thankfully, Ciseco provides the perfect solution – a wireless transceiver shield for the Arduino Uno, and an equivalent USB dongle

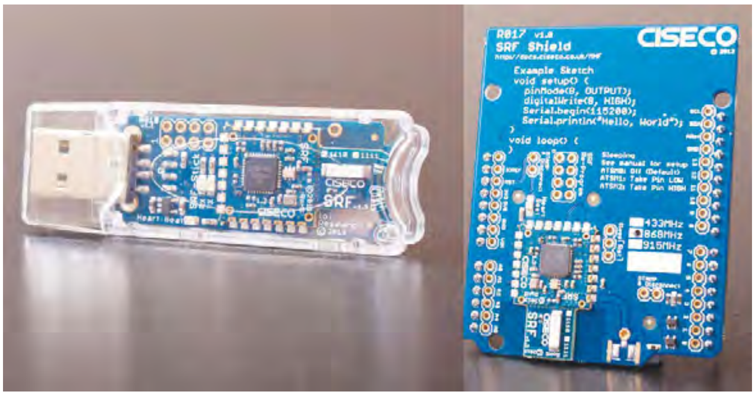


Fig.3. The core of the project – wireless link between the Arduino and the Raspberry Pi

that worked out of the box on the Pi. Communicating via the Arduino was simple – the code to do this was so straightforward it was printed on the silkscreen of the PCB. The boards are available from Cisesco UK directly, part numbers R010 and R017, at: **shop.ciseco.co.uk**. Bear in mind if you are interested in recreating this project that Ciseco have introduced a Pi-specific RF interface at half the cost of the USB interface, plus they have developed an Arduino Uno with an integrated RF interface.

Team building

After each team had pitched their ideas there was a brief period of settling in as people wandered around, spoke to different teams and then settled down into groups. We were lucky to assemble quite a large team – eight people – with a diverse range of skills. Most teams were much smaller, but we were fortunate that there was a lot of work

Fig.4. The winning team – although in the end, everyone was a winner

that could be split widely, and we ended up with an electronics expert (Tanya) an Arduino expert (Eoin) Raspberry Pi/Linux Sound expert (Harry) and the rest of us chipped in with sensor manufacture, testing and generally larking about. Having Harry join us was fortuitous because he had a strong background in Linux sound interfaces, and was able to integrate the ALSA sound drivers into the Raspberry Pi quicky, and to great effect. The result was a drum sound mixer and playback system that was superior to my dsPIC implementation, realised in a few hours rather than a few months of effort. Sure, my system was slightly cheaper and drew far less power, but for the intended setup, which would be mains powered, the cost and energy saving were irrelevant. By midnight we had essentially finished and were demonstrating (showing off) the system to the other teams.

(We have released our software as open source, and it can be downloaded from: **github.com/harryhaaren/Drum-Solo-Pants**. If there is interest, I will publish the hardware setup online. Drop us a message on the *EPE* forum: **www.chatzones.co.uk/discus**)

Sunday was a more relaxed day (for some) and gave us the chance to help out other teams. There were some really great project ideas, some too ambitious for a weekend, some daft, some downright silly (ours) but everyone had fun, and SciHackDay is now a date firmly fixed in our calendar.

My thanks to David McKeown for inviting me along, and to Jason, Eoin, Tanya, Becky, Richard, Michal and Harry for being such a great team to work with. I'm really looking forward to next year!

If you live in or near a major city you can expect to find a hack event of some description or other going on at some point in the year. SciHackDay was enormous fun, not just from the creativity side, but from sharing the experience and laughs with a group of like-minded individuals, many of whom have now become firm friends. My yearly calendar now rotates around the Maker fair in July, and the SciHackDay event in November.

Dublin SciHackDay happens every year around November time. To find out when the next event will be held, keep an eye on the website at **sciencehackdaydublin.com**. To find out what hackathons may be operating in your area post a message to the *EPE* web forum at: **www.chatzones.co.uk/discus**, where someone is bound to point you in the right direction.

Next month

In the next issue we return to our development platform and look at adding a newer, larger, low-cost LCD display, discovered by accident by one of our readers. It provides a full colour 2.2-inch panel with a QVGA resolution of 240 × 320 pixels, all accessible via an I^2C interface. The whole thing costs under £4, delivered. So it's ideal for an ultra-low-cost virtual reality display, or perhaps a simple oscilloscope. Find out all about it in next month's column!

Not all of Mike's technology tinkering and discussion make it to print. You can follow the rest of it on Twitter at: **@MikeHibbett**, *and from his blog at:* **mjhdesigns.com**

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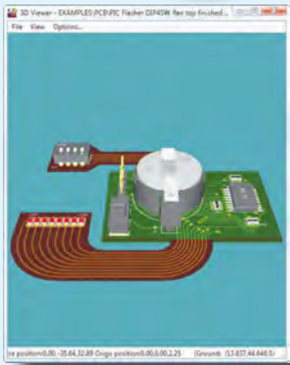

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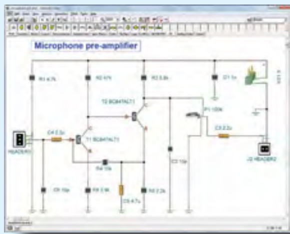

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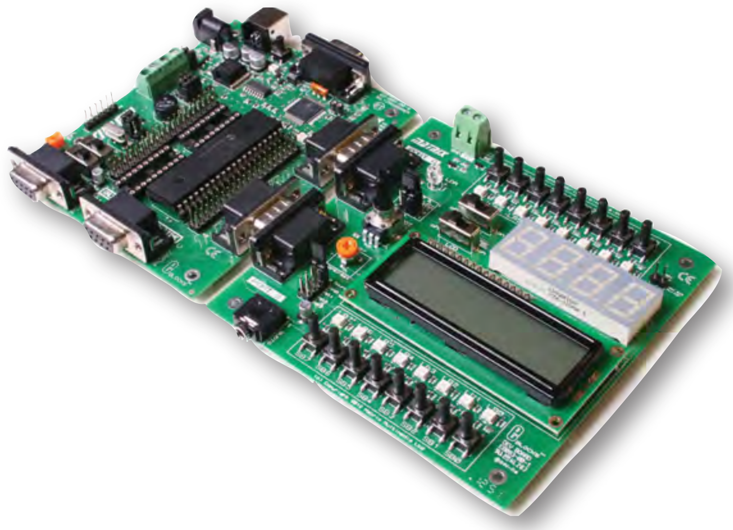

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SOFTWARE

ASSEMBLY FOR PICmicro V5

(Formerly PICtutor)

Assembly for PICmicro microcontrollers V3.0 (previously known as PICtutor) by John Becker contains a complete course in programming the PIC16F84 PICmicro microcontroller from Arizona Microchip. It starts with fundamental concepts and extends up to complex programs including watchdog timers, interrupts and sleep modes.

The CD makes use of the latest simulation techniques which provide a superb tool for learning: the Virtual PICmicro microcontroller, this is a simulation tool that allows users to write and execute MPASM assembler code for the PIC16F84 microcontroller on-screen. Using this you can actually see what happens inside the PICmicro MCU as each instruction is executed, which enhances understanding.

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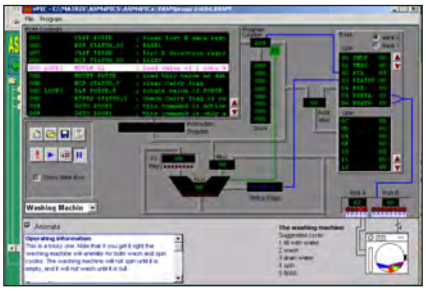

'C' FOR 16 Series PICmicro Version 5

The C for PICmicro microcontrollers CD-ROM is designed for students and professionals who need to learn how to program embedded microcontrollers in C. The CD-ROM contains a course as well as all the software tools needed to create Hex code for a wide range of PICmicro devices – including a full C compiler for a wide range of PICmicro devices.

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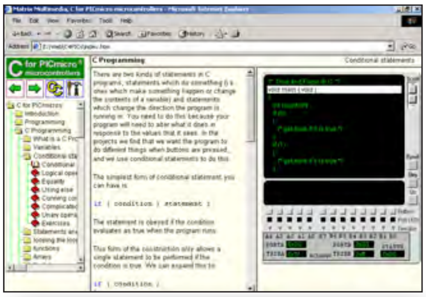

Minimum system requirements for these items: Pentium PC running, 2000, ME, XP; CD-ROM drive; 64MB RAM; 10MB hard disk space.
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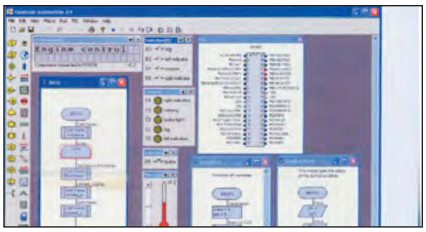
Please note: Due to popular demand, Flowcode PICmicro, AVR, DSPIC, PIC24 & ARM V6 are now available as a download. Please include your email address and a username (of your choice) on your order. A unique download code will then be emailed to you. If you require the CDROM as a back-up then please add an extra £14 to the price.



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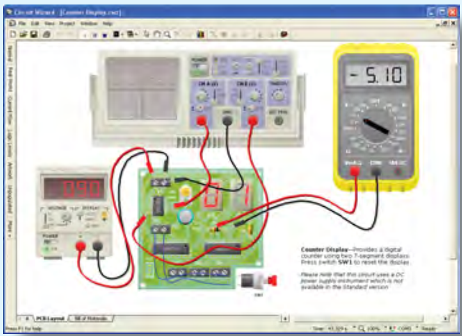

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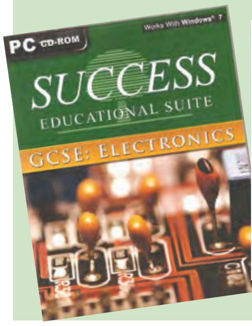

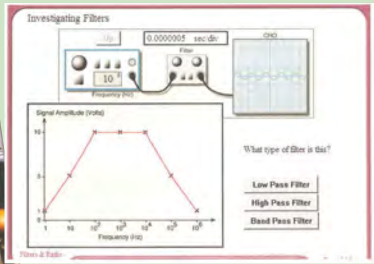

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Max's Cool Beans

By Max The Magnificent

Mastering meters – Part 3

I mentioned earlier in this mini-series that I like using analogue meters for my hobby projects because they offer a certain sense of style. In Part 1, we started by discussing the use of new, off-the-shelf analogue meters. In Part 2, we moved on to consider using antique analogue meters that one might find at 'Hamfests' and electronic flea markets. We also discussed how the first thing I do with these antique meters is to remove any internal (and possibly external) series and shunt resistors. This means that all we are left with is the resistance of the meter's coil.

The problem here is that we don't know what the resistance of the meter's coil is, and we can't use a multimeter to measure it, because doing so might actually blow the coil out of existence. Now, there is a way to measure the resistance of the coil without damaging it, and I will discuss this in my next column, but – as we shall see – we can actually get by without ever knowing this value.

As we've previously noted, the microcontroller unit (MCU) I'm using for the purposes of these discussions is an Arduino Uno, which boasts a 5V power supply. Also, we are using a pulse-width modulated (PWM) output pin to drive our meter (I'm using pin 11 on my Uno), where this pin can be assigned an 8-bit value from 0 to 255.

A more sophisticated circuit

In earlier columns, I showed the meter being driven directly from the MCU's output pin. Although this will work OK for most meters, I'm 'old school' and I prefer to protect my MCU's outputs. When you are driving inductive loads such as relays (solenoids) or motors, the back-EMF (electromotive force) can damage MCU pins.

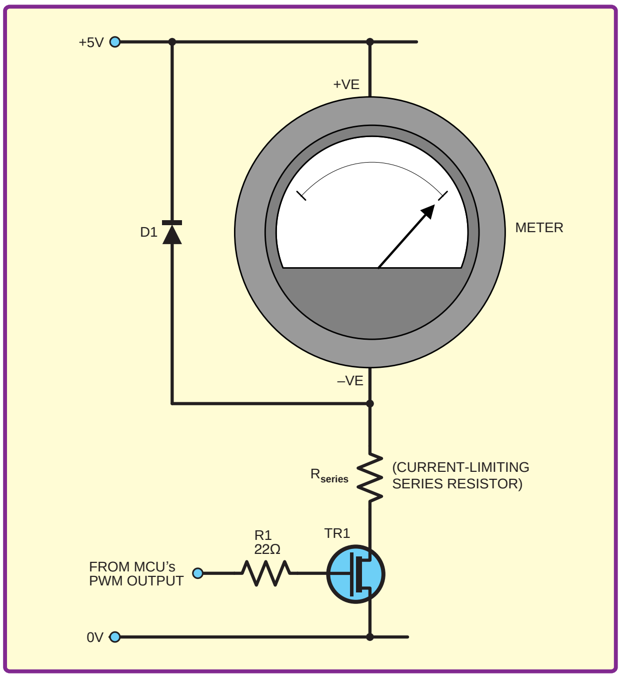

Fig.1. Circuit used to drive the analogue meter

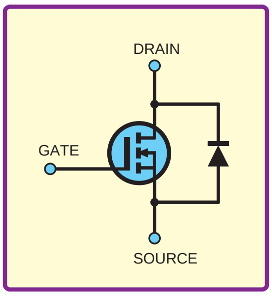

Fig.2. The BS170 contains a protective diode

Similarly, some analogue meters have the potential (no pun intended) to damage the MCU outputs driving them, and once you've blown a pin inside your MCU there's no going back. Hence, I prefer to use the circuit shown in Fig.1.

One point worth noting is that you very often see the R_{series} current-limiting resistor presented 'before' the meter; ie, between the 5V rail and the meter's +ve terminal in this sort of diagram. In Fig.1, however, I've shown R_{series} as appearing 'after' the meter – between the meter's negative (–ve) terminal and the rest of the circuit. In reality, it makes no difference to the functioning of the circuit whether this resistor appears before or after the meter. The reason I show R_{series} as being after the meter is this makes things easier to wire up when we come to the real-world use of the meter.

Diode D1 is used to protect the MCU pin against back-EMF. I'm using a general-purpose 1N4007 (1000V, 1A) device for this purpose. Transistor TR1 is a general-purpose N-channel enhancement mode MOSFET. I'm using a BS170 (60V 500MA) in a TO-92 package, because it's easy to obtain – and a cheap-and-cheerful device. The reason for using a field-effect transistor (FET) instead of a bipolar junction transistor (BJT) is that there is almost negligible voltage drop across the FET when it is active, as opposed to the ~0.7V voltage drop across a BJT. Ideally, we'd also have a diode strapped across TR1's source and drain terminals to protect it from back EMF; and happily the BS170 already contains such a diode inside its package, as illustrated in Fig.2.

Determining the R_{series} value

So, how are we going to set about determining what value of R_{series} to use? Based on the meters I've been playing with for my own projects, this can range anywhere from 370Ω (or sometimes even lower) to 92kΩ (or higher). In order to evaluate this on a meter-by-meter basis, I created the 'Series Resistance Test Unit' shown in Fig.3.

All this really comprises is four linear potentiometers connected in series. In the front row we have a 1kΩ pot. In the back row, from left-to-right, we have 10kΩ, 100kΩ, and 500kΩ pots. These are all wired such that turning them fully clockwise sets them to their maximum value. Coming in from the left we have +5V from the Arduino (the red wire) and the signal from the drain of Transistor TR1 (the black wire). Going out from the right we have the red and green wires that will be connected to the positive (+ve) and –ve terminals on the meter, respectively.

Located in the front-middle we have a 2-pole, 3-throw switch. When I built this, I wasn't sure if all analogue meters had their +ve terminals on the same side, so I included this switch to allow me to quickly and easily

Fig.3. Max's super-duper Series Resistance Test Unit

reverse the polarity of the outputs. Since then, I've been told that all analogue meters have their +ve terminal on the left-hand side (when looking from the back), and this has certainly been true of all the meters I've looked at myself, so I probably wouldn't bother including this switch if I were to build another version of this unit. Having said this, I would still include a 2-pole, 2-throw switch to allow me to disable the outputs if required.

Using the Series Resistance Test Unit

1. The first thing I do is to set the switch to its center position (disabling the outputs) and turn all of the pots to their maximum values.
2. Next, I power up the Arduino and load a simple program whose only purpose is to drive a maximum value of 255 to the PWM output connected to the input of R1, thereby turning TR1 full on.
3. I then turn the switch to its right-hand position to enable the outputs.
4. Next, I start turning the 500kΩ potentiometer slowly anticlockwise while keeping an eagle eye on the meter. For all the meters I've tested thus far, I end up with the 500kΩ pot at its minimum value, but you never know what the next meter will bring.
5. Then I start turning the 100kΩ pot slowly anticlockwise. Let's assume that this also reaches its minimum value with no effect, in which case I start turning the 10kΩ pot slowly anticlockwise.
6. For the purpose of these discussions, let's assume that the meter's needle now starts to rise (move from left-to-right). I keep turning the 10kΩ pot slowly anticlockwise until the needle approaches its full-scale deflection (FSD)
7. At this point I switch to the 1kΩ pot to make the final fine adjustment to get the needle to its FSD.
8. Now I disconnect the meter and the Arduino, and I use my multimeter to measure the current value of my potentiometer chain. This is going to be my R_{series} value for this meter.

The real R_{series} implementation

Of course, this R_{series} value may be a non-standard value. And even if it is a standard value, all resistors have tolerance associated with them, so a 10kΩ resistor with a 5% tolerance may actually be anywhere between 9.5kΩ and 10.5kΩ. Last, but not least, there may be minor differences between various values in this test circuit and in my final implementation circuit. For all these reasons, the current-limiting resistor I actually use on my Arduino's prototyping board is formed with a fixed resistor and a multi-turn (I prefer 25 turns) trimmer pot connected in series. Both of these will be selected to have a value as close as possible to 2/3 of the R_{series} value we just calculated/measured.

So, now we are ready to make the final adjustments. We start by setting the trimmer pot to its maximum resistance, which means we now have an R_{series} value of 4/3 R_{series}. Once again, we power up the Arduino and set its PWM output to be driving a maximum value of 255, in which case we expect to see the meter's needle substantially below its FSD. Now we slowly turn the trimmer pot until the needle just makes its FSD.

But wait, there's more!

We still have some more work to do, including creating new faceplates for our meters that better reflect their new functions. In the case of my Vetinari Clock project for example, I wanted my main meter to display unit of hours, not its original units of resistance. Related considerations include the linearity of the meters' movements and how we correct for non-linear movements. Fear not, because this is not too complicated, and all will be revealed in the next – and final – segment of this meter miniseries.

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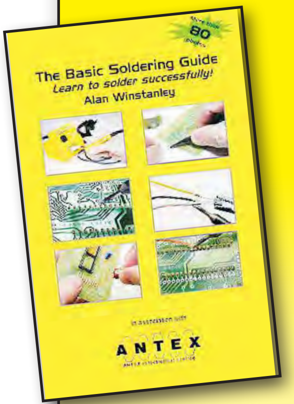

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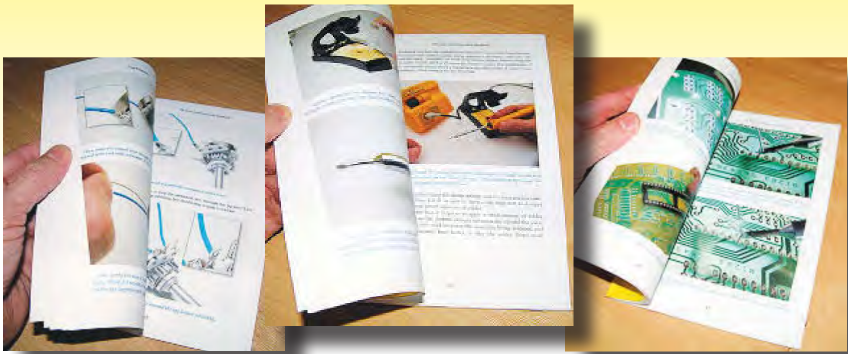

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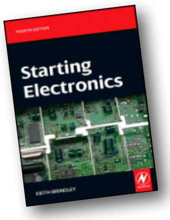

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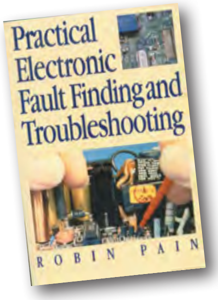

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All prices include VAT and postage and packing. Add £2 per board for airmail outside of Europe. Remittances should be sent to **The PCB Service, Everyday Practical Electronics, Wimborne Publishing Ltd., 113 Lynwood Drive, Merley, Wimborne, Dorset BH21 1UU. Tel: 01202 880299; Fax 01202 843233; Email: orders@epemag.wimborne.co.uk. On-line Shop: www.epemag.com**. Cheques should be crossed and made payable to *Everyday Practical Electronics* **(Payment in £ sterling only)**.

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★ All software programs for *EPE* Projects marked with a star, and others previously published can be downloaded free from the Library on our website, accessible via our home page at: **www.epemag.com**

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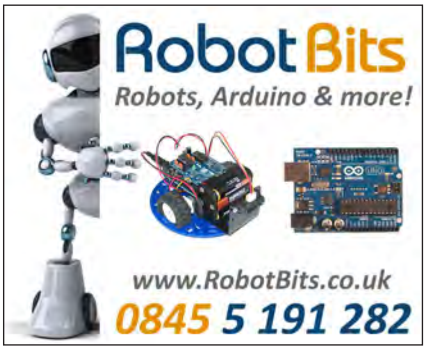

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Next Month

A Rubidium Frequency Standard for a Song

How would you like to have a precision rubidium frequency standard on your workbench to enable you to make laboratory-standard frequency measurements? It's now possible and for a very low price – just buy a used Rb-vapour frequency standard on-line and build a simple power supply and buffer circuit.

10A/230V Speed Controller for Universal Motors – Part 2

Last month, we described the features and circuit details of our new *Speed Controller*. This month we move on to construction and troubleshooting.

USB/RS-232C Interface

Want to connect an older test instrument or PC peripheral fitted with a 'legacy' serial RS-232C interface to your late-model PC or laptop? That is a real problem with today's PCs, which only provide USB ports. Here is the solution: build this very small *USB to RS-232C Interface*. We're sure you'll find no end of uses for it!

Teach-In 2015 – Part 3

In April's *Teach-In 2015*, we feature the design and construction of a simple headphone amplifier, and introduce you to power amplifiers, explaining classes of operation. Plus, just for good measure, we'll show you how to use Circuit Wizard to design your own printed circuit boards.

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